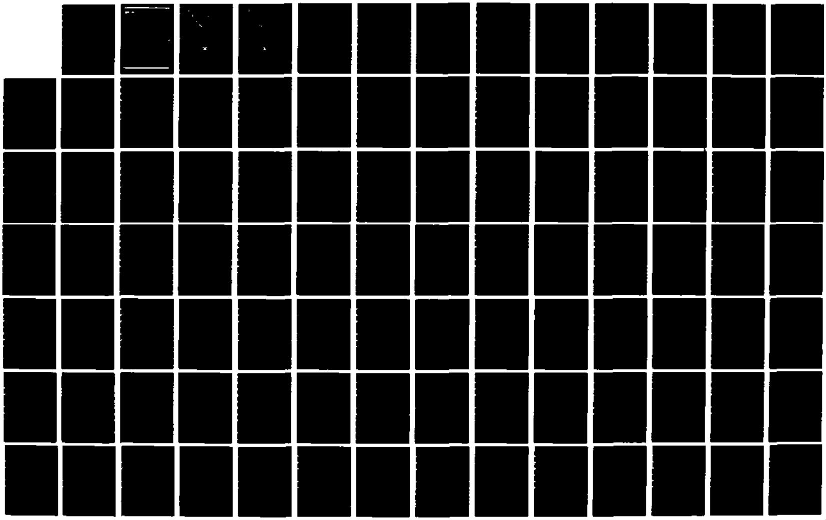
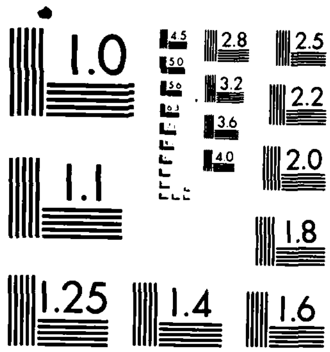
AD-A162 116 DSCS (DEFENSE SATELLITE COMMUNICATIONS SYSTEM)/WIDEBAND 1/2
SHF (SUPER HIGH F (U) M/A-COM GOVERNMENT SYSTEMS DIV
VIENNA VA L RICARDI 29 NOV 85 DCA/MSO-85-3-PT-2
UNCLASSIFIED DCA100-84-C-0009 F/G 17/2 NL






MICROCOPY RESOLUTION TEST CHART
NATIONAL BUREAU OF STANDARDS-1963-A

(2)

M/A-COM

GOVERNMENT SYSTEMS DIVISION

8619 WESTWOOD CENTER DRIVE
VIENNA, VA 22180

M/A-COM LINKABIT, Inc.
Log # MSO-85-114a
Copy # 9

AD-A162 116

# DSCS/Wideband SHF Enhancements and EHF Antenna Recommendations for the Mid-(1986-1991) and Far-Term (1992-2000) Periods

Part II

Final Report

Task MSO85-3



November 1985

Prepared by M/A-COM LINKABIT, Inc.
Under Contract DCA100-84-C-0009

Submitted to
Defense Communications Agency
Center for Command and Control,
and Communications Systems, Code A800
8th & S. Courthouse Road
Arlington, VA 22204





85 12 - 027

511258.0

# CENTER FOR COMMAND AND CONTROL, AND COMMUNICATIONS SYSTEMS ($C^4S$)

" EXCELLENCE IN $C^3$ SYSTEMS FOR NATIONAL DEFENSE"

## DSCS/Wideband SHF Enhancements and EHF Antenna Recommendations for the Mid- (1986-1991) and Far-Term (1992-2000) Periods

**Part II**

**Final Report**

**November 1985**

**DEFENSE COMMUNICATIONS AGENCY**

$C^4S$ ____________________

CY ________OF ________CYS

511258.0

# CENTER FOR COMMAND AND CONTROL, AND COMMUNICATIONS SYSTEMS ($C^4S$)

" EXCELLENCE IN $C^3$ SYSTEMS FOR NATIONAL DEFENSE"

## DSCS/Wideband SHF Enhancements and EHF Antenna Recommendations for the Mid- (1986-1991) and Far-Term (1992-2000) Periods

**Part II**

**Final Report**

**November 1985**

## DEFENSE COMMUNICATIONS AGENCY

$C^4S$ ____________________

CY ________ OF ________ CYS

511258.0

CONTENTS

CONTENTS (Cont'd)

LIST OF FIGURES

LIST OF TABLES

## ACRONYMS AND ABBREVIATIONS

| | |
|---|---|
| BER | Bit Error Rate |
| BFN | Beam-Forming Network |
| CDMA | Code Division Multiple Access |
| COMSAT | Communication Satellite |
| CP | Circular Polarization |
| CW | Continuous Wave |
| DAMA | Demand Assigned Multiple Access |
| DCEC | Defense Communication Engineering Center |
| DOF | Degrees of Freedom |
| DMUX | Demultiplex |
| DPS | Dual Polarized System |
| DSCS III | Defense System Communication Satellite |
| DSCSOC | DSCS Operations Center |
| EHF | Extremely High Frequency |
| EIRP | Effective Isotropic Radiated Power |
| ESAAP | EHF Satellite Adaptive Array Program |
| FDMA | Frequency Division Multiple Access |
| FFJ | Frequency Follower Jammer |
| FH | Frequency Hopping |
| FOV | Field of View |
| G/T | Gain-To-Noise Temperature |
| GDA | Gimbaled Dish Antenna |
| HPA | High Power Amplifier |
| HPBW | Half Power Beamwidth |
| IF | Intermediate Frequency |
| JCS | Joint Chiefs of Staff |
| JLE | Jammer Locator Equipment |
| LHCP | Left Hand Circularly Polarized |
| LNA | Low Noise Amplifier |
| MBA | Multiple Beam Antenna |
| MUX | Multiplex |
| NSSK | North-South Stationkeeping |
| PN | Pseudonoise |
| QPSK | Quadrature Phase Shift Keying |
| RADC | Rome Air Development Center |
| RF | Radio frequency |
| RHCP | Right Hand Circularly Polarized |
| RMS | Root Mean Square |
| S/N | Signal-to-Noise Ratio-overall |
| SATCOM | Satellite Communication |
| SHF | Super High Frequency |
| TDM | Time Division Multiplexer |
| TDMA | Time division multiple access |
| TPA | Thinned Phased Array |
| TWTA | Traveling Wave Tube Amplifier |
| VPD | Variable Power Divider |

# CHAPTER 1
# EXECUTIVE SUMMARY

This report considers many facets of adding dual polarization capability to the Defense Satellite Communications System Phase III (DSCS III) super high frequency (SHF) antenna subsystem and the design features of a beam-switching antenna for an extremely high frequency (EHF) communication package on DSCS III, or a follow-on satellite. Wideband user scenarios are emphasized throughout but lower-data-rate user requirements are also discussed. The report summarizes a brief study of the salient issues and provides quantitative assessment on several of the key issues. This section gives a brief overview of the major issues and a general understanding of the results of the study. Recommended further action or studies are listed at the end of this section.

## 1.1 DUAL-POLARIZED DSCS III SHF ANTENNA SYSTEM

Operational success of the DSCS III and development of the related terminal community has resulted in a need for an increase in the payload's communication capacity. Currently the up and downlinks are configured to provide frequency division multiple access (FDMA). The effective isotropic radiated power (EIRP) produced on the multiple downlinks and the gain-to-noise temperature (G/T) existing on the multiple uplinks are large enough to accommodate increased communication if the associated terminals have sufficient EIRP and G/T. However, DSCS III will be bandwidth limited if the terminals have sufficient EIRP on their uplink and G/T on their downlink. Since the total bandwidth of the DSCS III payload is 375 MHz*, it cannot have a communications capacity greater than about 375 Mbps given rate one-half coded quaternary modulation

*Usable bandwidth will increase to 405 MHz starting with DSCS III B8.

(e.g., quadriphased shift keying (QPSK)). The need to assign channels to communities that use FDMA can result in some channels being under utilized; this may, in a practical sense, reduce the maximum capacity to about 200 Mbps. Significant increases in terminal or payload EIRP or G/T in conjunction with higher-order modulation formats and higher code rates could be used to increase the throughput communication capacity. However, both the antijam and unprotected communication capacity may be doubled through use of a dual-polarized antenna system. This results from the effective increase in bandwidth through frequency reuse possible with independent transmission on each of two orthogonally polarized radio waves. This study report addresses the design and salient performance characteristics of a dual-polarized antenna system.

### 1.1.1 Current Antenna System

The current DSCS III SHF payload is schematically summarized in Figure 1-1. It consists of six channels varying in bandwidth from 60 to 85 MHz. Its uplink antenna suite consists of a 61-beam multiple beam antenna (MBA) and two earth coverage antennas. Five antennas, consisting of two earth coverage antennas, two MBAs and a gimballed dish antenna (GDA) serve the downlink. Redundancy switching permits the channels to be connected to most any combination of up and downlink antenna types. The beam of the uplink MBA can be phase and amplitude weighted to produce arbitrary shaped radiation patterns with a null toward interference sources located in the antenna's field of view (FOV). The downlink MBAs can also shape their radiation patterns. There is no need for them to produce a null in the direction of an interference source; consequently, the downlink MBA's beam are only amplitude weighted.

Jammer locater equipment (JLE) attached to the uplink MBA telemeters signal location information to a DSCS Operations

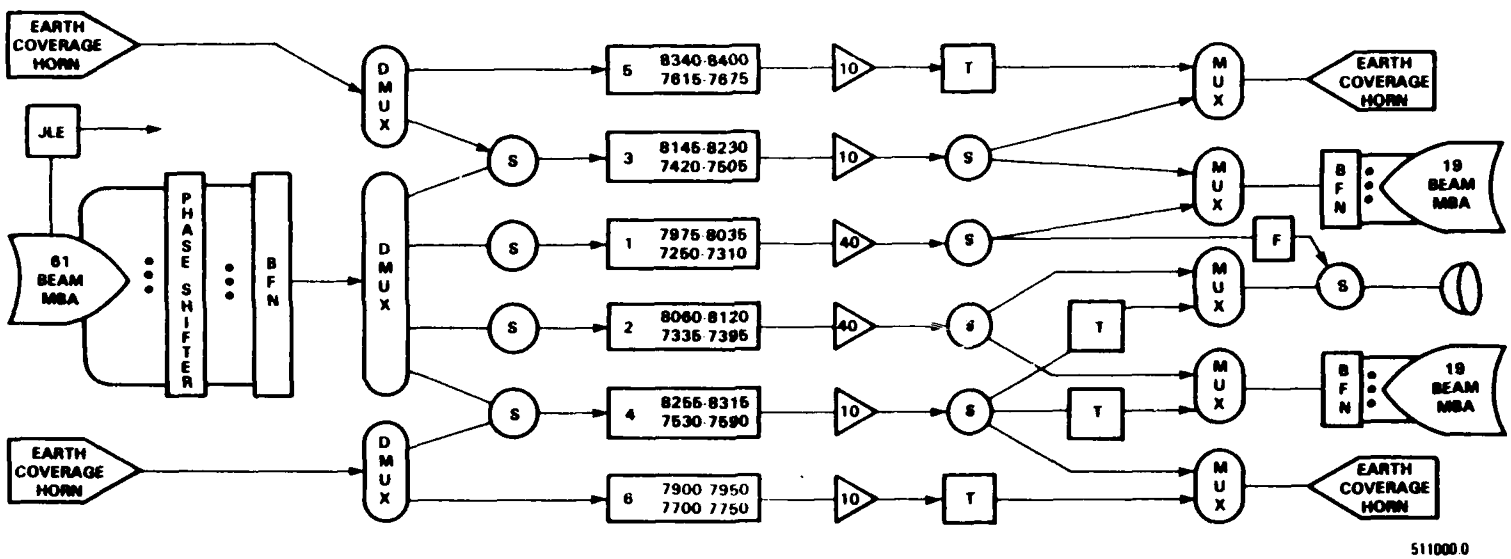



Figure 1-1. DSCS III Payload*

*Note Channel 1-6 Frequency Plan is for Payloads B-7 and Earlier

center (DSCSOC). The DSCSOC processes this information to determine the location of jammers. Using the jammer-location data, the known-beam radiation patterns of DSCS III, and the desired jam-free, or quiescent, radiation pattern, an optimum set of antenna weights is computed. The appropriate information is transmitted to the spacecraft and the weights are installed. This reconfiguration process can take from one to several minutes and results in jammer signal suppression through spatial antijam discrimination.

### 1.1.2 Upgrade Antenna

The uplink MBA has dual-polarized ports, right-hand circularly polarized (RHCP) and left-hand circularly polarized (LHCP), on all beams; the RHCP ports are active and the LHCP ports are currently terminated in a matched load. These loads can be removed and a duplicate of the current beam-forming network installed as indicated in Figure 1-2. Similar modification of the downlink MBAs, GDA, and earth coverage horns is also possible. Figure 1-2 shows a recommended strawman channelization using LHCP signal ports that would provide one 50-MHz and two 155-MHz channels. The additional crosspolarized (LHCP uplink) channels are in principle isolated from the current, or copolarized, channels. If the terminals are modified accordingly, the throughput capacity of DSCS III, determined on virtually any basis, could be doubled. It may be necessary to install separate JLE on the crosspolarized channels; further study is required.

Notice that multiplexing functions, switches, and traveling wave tube amplifiers (TWTAs) are also indicated in Figure 1-2 to point out the need for these devices and indicate how they could be incorporated into a dual-polarized system.

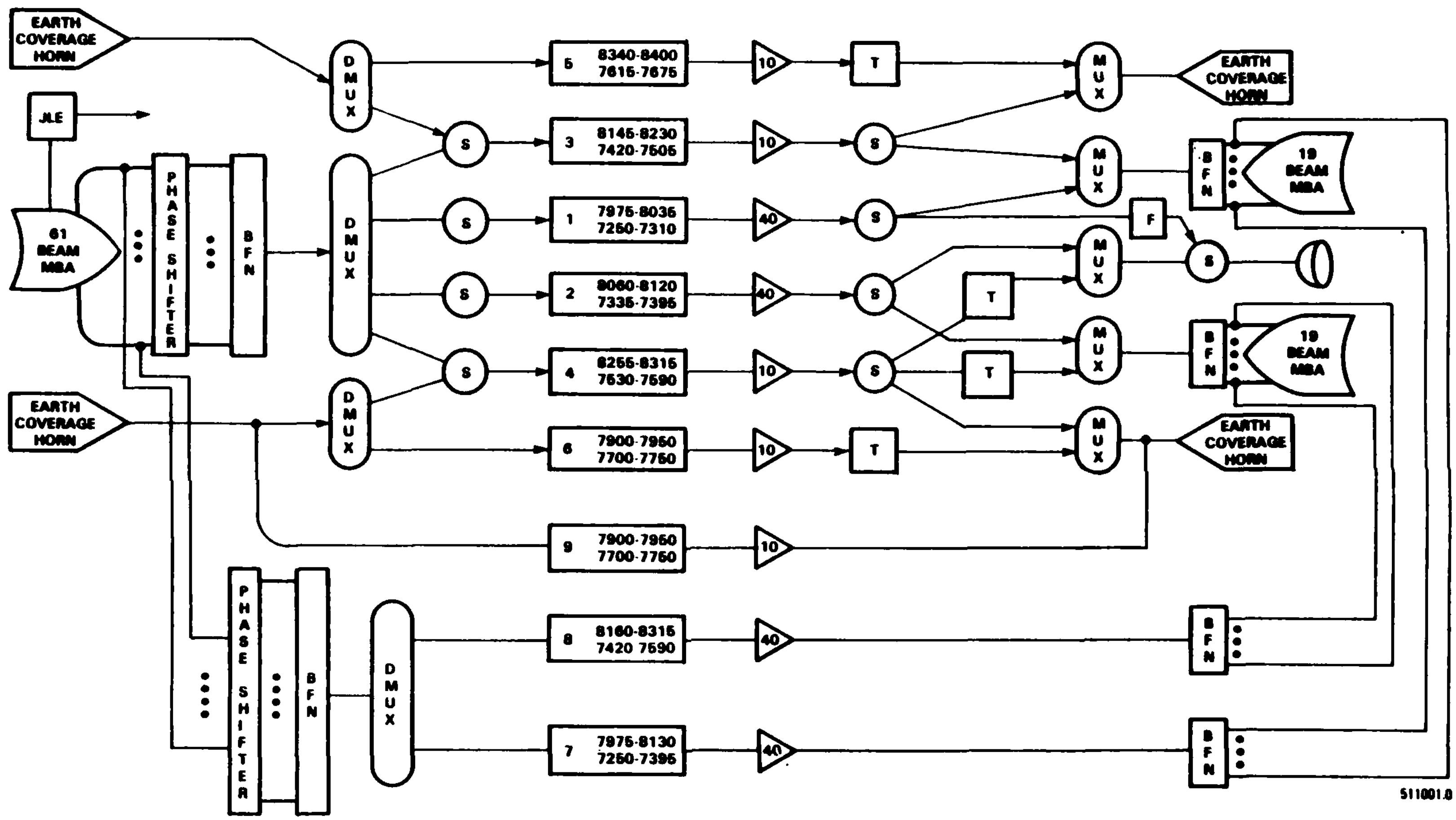



Figure 1-2. DSCS III Dual-Polarized Payload*

*Note Channel 1-6 Frequency Plan is for Payloads B-7 and Earlier

### 1.1.3 General Performance Requirements

Installation of an autonomous nulling algorithm on board the spacecraft would make it possible for the satellite to "continuously" adapt the uplink antenna, maintaining the jammer in a null regardless of the spacecraft attitude variation. In particular, this would relax the requirement for North-South stationkeeping (NSSK) and save about 400 lbs of fuel necessary to carry out this function. This savings should be confirmed by further study and, if applicable, used to provide the increased communication payload weight necessary to support a frequency reuse upgrade to DSCS III and also provide an on-board autonomous nulling algorithm that would improve the nulling performance of the DSCS III uplink MBA.

Unfortunately, dual-polarized channels are vulnerable to self jamming if purity of polarization is not maintained within specified standards. The required purity of polarization is difficult to achieve for two major reasons:

1. A circularly polarized wave will undergo a change in polarization when it propagates through rain or ice.

2. The uplink and downlink MBAs cannot be made to have identical copolarized or dual-polarized performance characteristics.

If these imperfections can be limited such that the polarized waves radiated or received by the payload's antenna have an axial ratio less than 1 dB, dual-polarization channels will operate satisfactorily. Satisfactory operation may not be possible in those scenarios where a terminal with a very large EIRP operates at the same frequency (but different polarization) as a low-EIRP terminal.

### 1.1.4 Terminal Considerations

Currently the DSCS terminals transmit RHCP and receive LHCP signals. Their radio frequency (RF) and antenna system would have to be modified to permit transmission on LHCP and reception on RHCP. It may be desirable to modify the terminals so they transmit and receive either sense of circular polarization (CP). Imperfection in the satellite's transmitted "circularly" polarized wave and the depolarization of the wave when it propagates through rain can be compensated by installing a polarization tracker in the terminal's antenna (see Figure 1-3). Using the measured polarization of the received signal the terminal could correct for the rain-depolarization effect by transmitting an appropriately polarized wave on the uplink. Satisfactory operation on dual-polarized channels may require this polarization tracking capability at each terminal. It would be of no advantage to install a polarization tracker on the satellite since signals received simultaneously from different terminals will have a slightly different polarization and it would be impossible for the satellite to optimize reception of signals from all terminals at the same time.

### 1.1.5 System Considerations

If the copolarized channel is RHCP with zero dB axial ratio, the crosspolarized channel will couple to the copolarized channel if its axial ratio is greater than zero dB. The degree of coupling is given in Figure 1-4. Notice that a one dB axial ratio results in -25-dB coupling between channels. If the signal strengths in each channel are nearly equal, they probably will not jam one another. If the signals in one channel are 20 dB larger than those in the other channel, the stronger signal channel will jam the weaker channel--perhaps preventing communication via that channel.

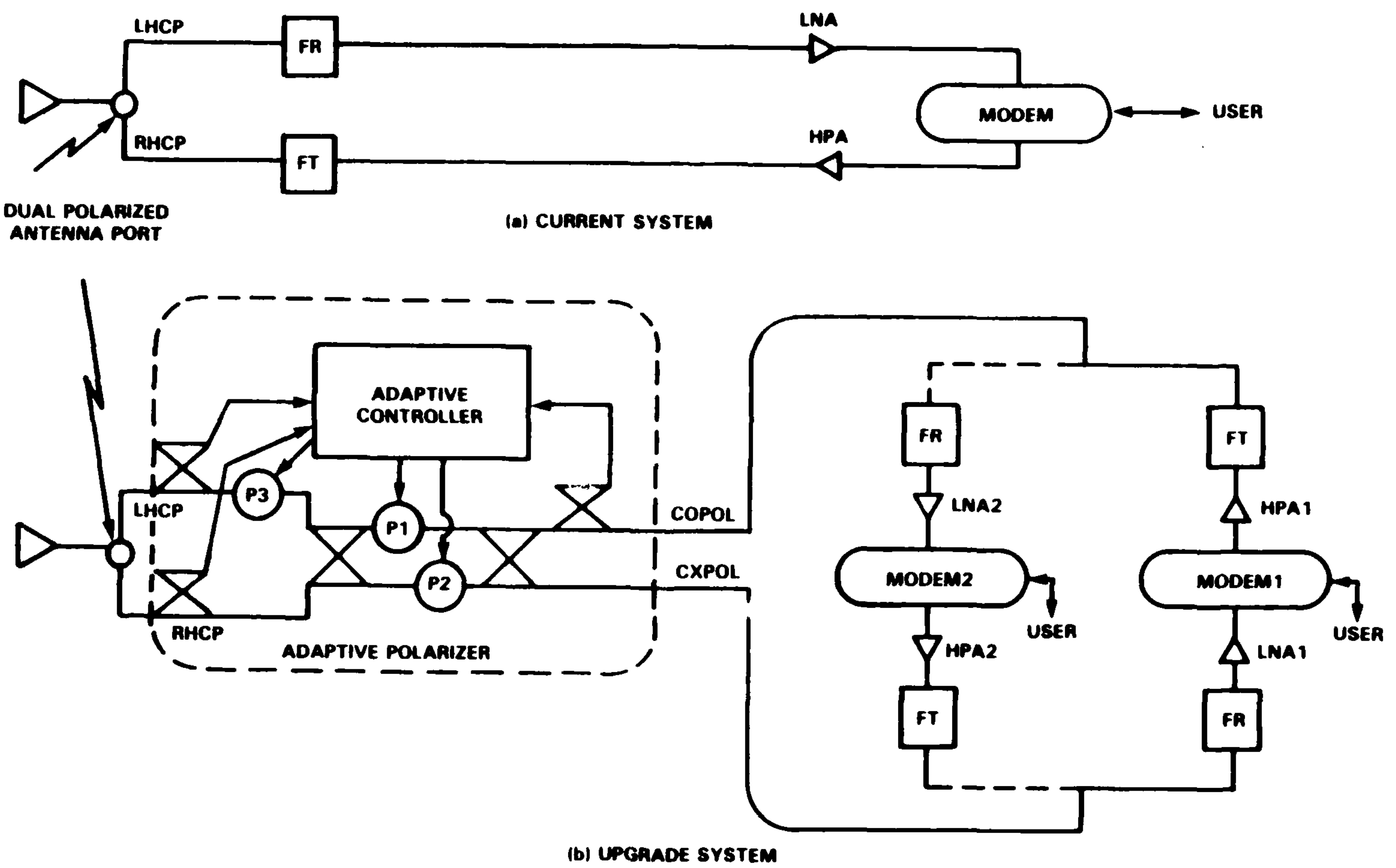



Figure 1-3. Typcial Terminal RF System

511002 0

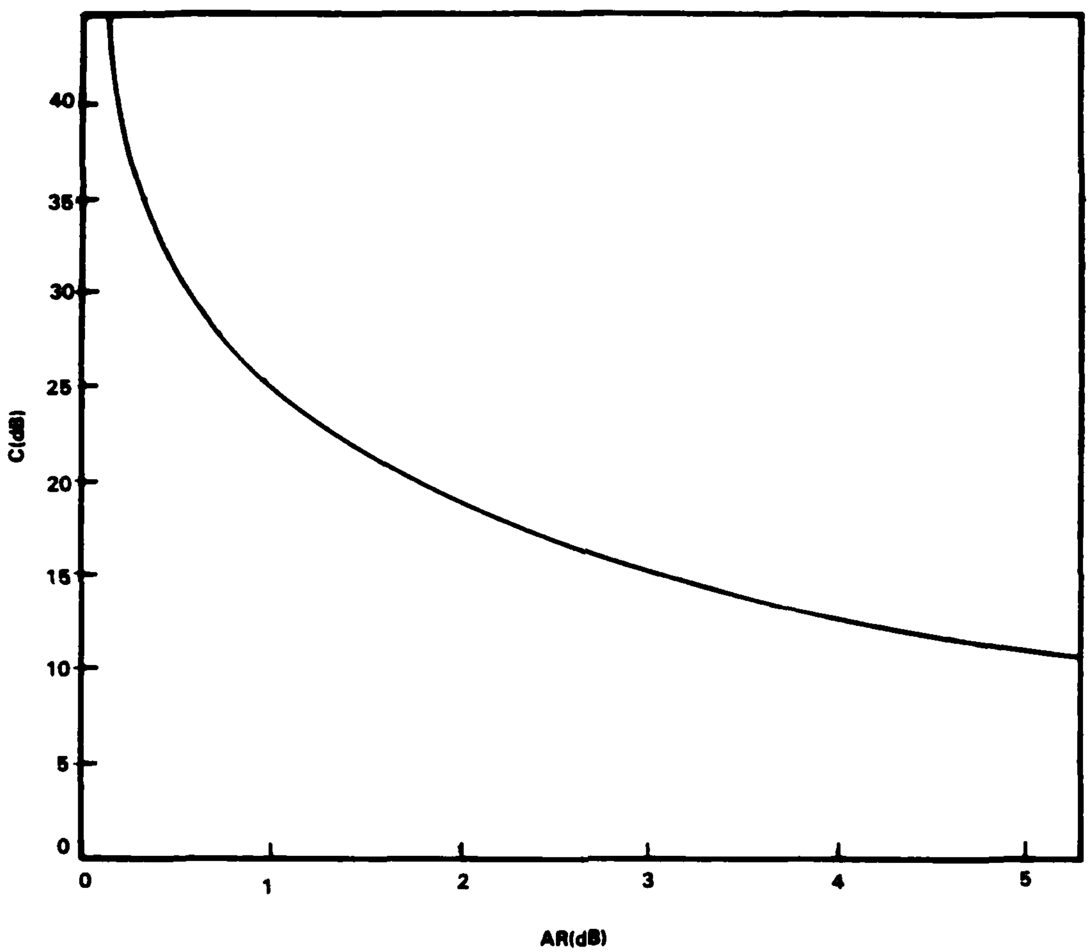



**Figure 1-4. Coupling Due to Cross Polarization**

511003.0

In the foregoing example a one-dB axial ratio was assumed. It will be difficult for both the spacecraft and the terminal to have an axial ratio small enough to guarantee -25 dB coupling in a rain-free environment. Moderate-to-heavy rain storms can reduce coupling to -15 dB, unless a polarization tracker (Figure 1-3) is used. The terminal and spacecraft axial ratios may be required to be as low as 0.3 dB to tolerate the expected depolarization due to rain. To achieve a 0.3-dB axial ratio would be a significant technical challenge for both the MBA lens used on DSCS III and the terminal segment. Further study is required to assess this interactive effect accurately and develop a final specification on the axial ratio of a dual-polarized DSCS MBA.

## 1.2 ANTENNA DESIGN OF A WIDEBAND EHF PACKAGE

Studies have shown that EHF Military Satellite Communication (MILSATCOM) systems are potentially more robust than lower-frequency MILSATCOM systems due to increased spatial discrimination and reduced disparity between expected jammer and user-terminal EIRP. (Note that technology predictions show a $f^{-2}$ RF power trend for jammers.) Increased bandwidth also provides a potentially larger communication capacity than at SHF or lower-frequency MILSATCOM systems. Because of this and the shortage of bandwidth and communication capacity at SHF, many studies have been conducted to determine the characteristics of an EHF antenna for a follow-on DSCS III payload.

These studies have chosen essentially two fundamental antenna designs: the switched-beam MBA and the thinned phased array (TPA). Characteristically the TPA has better nulling resolution than the MBA; that is, the TPA can overcome a smaller jammer-user separation than the MBA. On the other hand, the TPA can be completely disabled by a group of small jammers located thousands of miles from a user; the MBA handles

this scenario very well. Since sanctuary jammers are potentially a more realistic threat than a near-in jammer, it is unwise to consider the TPA as an EHF-survivable, jam-resistant antenna. Consequently, a switched-beam MBA is proposed as a candidate EHF antenna. The specific 285-beam MBA proposed here has many desirable properties; however, the exact number of beams required and the desired aperture size are legitimate areas for future detailed tradeoff studies. Figure 1-5 shows the candidate EHF antenna system.

### 1.2.1 Description

The candidate EHF antenna consists of a 24" diameter dielectric lens illuminated by a 285-horn feed system. An array of 16 switching trees connects 16 of the 285 feed horns (beams) to a 16-degree freedom nulling algorithm. The array of 285 beams covers, contiguously, the earth FOV. Beams can be selected to give point or area coverage in less than one microsecond. There can be one or more narrow beams, one or more area beams, or a mixture of up to 16 different beams simultaneously; although normal operation would require probably less than three beams to be formed simultaneously. That is, one beam each for high-, medium-, and low-data-rate users operating in time division multiple access (TDMA), FDMA, or TDMA/FDMA. A tabulation of various terminals sizes, data rates, and margins is presented in Table 1-1. The values in Table 1-1 assume the satellite's uplink antenna is generating a single high-gain beam and there is 12 dB of rain attenuation in addition to 2.34 dB of atmospheric attenuation. Wider area coverage beams, simultaneous formation of more than one beam, and a mixture of both, may reduce the uplink antenna gain and the indicated margin by as much as 9 dB. Note that data rate as high as 100 Mbps to a 108-dBW EIRP terminal can be supported with an excess margin of 26 dB in addition to a 14.4-dB margin for rain and atmospheric attenuation.

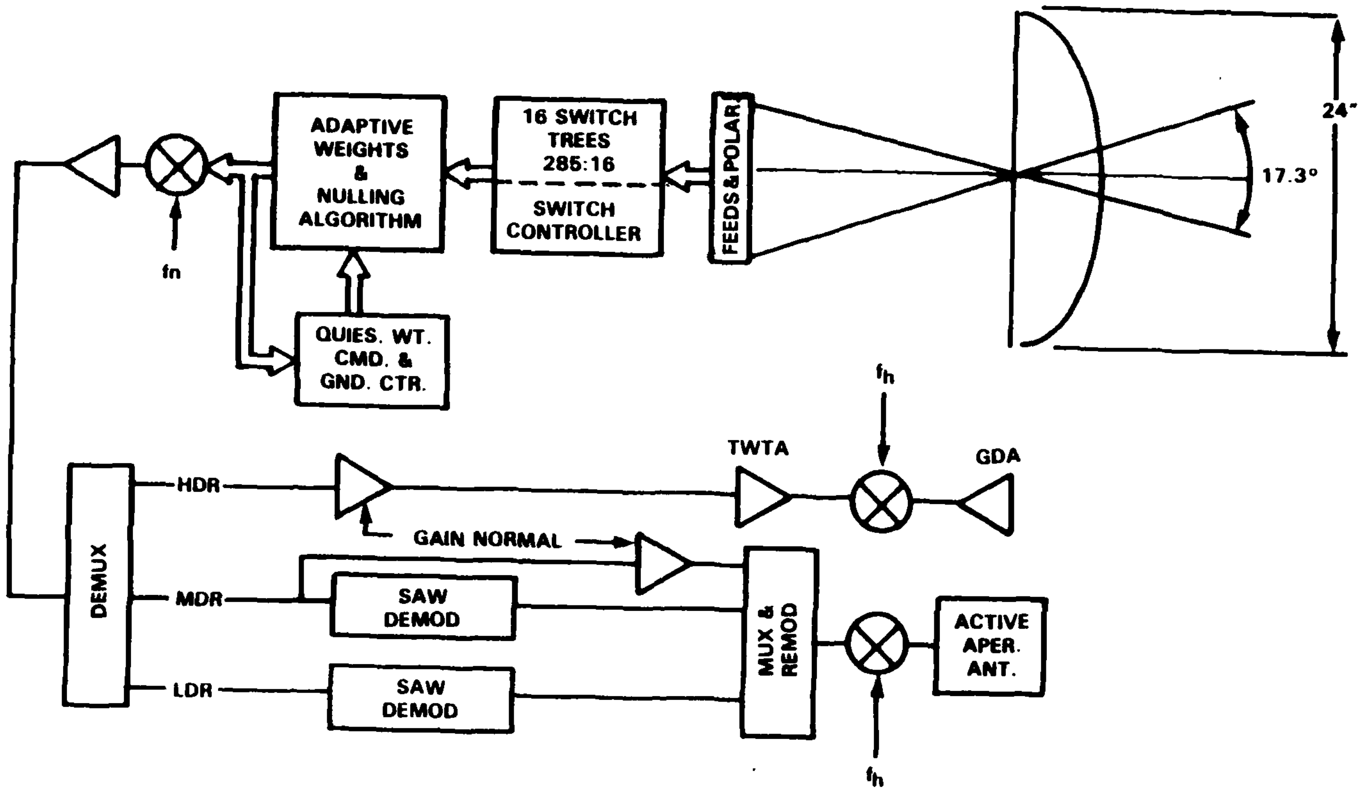



Figure 1-5. Candidate EHF Antenna System

511004.0

## Table 1-1. Link Budgets

| | | | | | | |
|---|---|---|---|---|---|---|
| **UPLINK BUDGET** | | | | | | |
| **TERMINAL CHARACTERISTICS** | | | | | | |
| FREQUENCY (GHz) | 44.50 | 44.50 | 44.50 | 44.50 | 44.50 | 44.50 |
| WAVELENGTH (IN) | .27 | .27 | .27 | .27 | .27 | .27 |
| ANTENNA APERATURE DIAMETER (FT) | 40.00 | 30.00 | 20.00 | 10.00 | 8.00 | 4.00 |
| HALF POWER BEAMWIDTH (DEG) | .04 | .05 | .08 | .15 | .19 | .39 |
| ANTENNA GAIN (dB) | 72.60 | 70.10 | 66.58 | 60.56 | 58.62 | 52.60 |
| RF CIRCUIT LOSSES (dB) | 1.50 | 1.50 | 1.50 | 1.50 | 1.50 | 1.50 |
| POINTING LOSS (dB) | .00 | .00 | .00 | .00 | .50 | .30 |
| TRANSMIT OUTPUT POWER (dBW) | 37.00 | 33.00 | 30.00 | 17.00 | 15.00 | 15.00 |
| EIRP (dBW) | 108.10 | 101.60 | 95.08 | 76.06 | 71.62 | 65.80 |
| **PROPAGATION CHARACTERISTICS** | | | | | | |
| SPREAD LOSS (SYN. SAT.) (dB) | 217.51 | 217.51 | 217.51 | 217.51 | 217.51 | 217.51 |
| ATMOSPHERIC ATTENUATION (dB) | 2.34 | 2.34 | 2.34 | 2.34 | 2.34 | 2.34 |
| RAIN ATTENUATION (dB) | 12.00 | 12.00 | 12.00 | 12.00 | 12.00 | 12.00 |
| TOTAL PROPAGATION LOSS (dB) | 231.85 | 231.85 | 231.85 | 231.85 | 231.85 | 231.85 |
| **SATELLITE CHARACTERISTICS** | | | | | | |
| ANTENNA APERTURE DIAMETER (IN) | 24.00 | 24.00 | 24.00 | 24.00 | 24.00 | 24.00 |
| HALF POWER BEAMWIDTH (DEG) | .77 | .77 | .77 | .77 | .77 | .77 |
| ANTENNA GAIN (dB) | 46.58 | 46.58 | 46.58 | 46.58 | 46.58 | 46.58 |
| RF CIRCUIT LOSS (dB) | 3.50 | 3.50 | 3.50 | 3.50 | 3.50 | 3.50 |
| POINTING LOSS (dB) | .00 | .00 | .00 | .00 | .00 | .00 |
| SYSTEM NOISE TEMPERATURE (°K) | 1500.00 | 1500.00 | 1500.00 | 1500.00 | 1500.00 | 1500.00 |
| NOISE FIGURE (dB) | 7.99 | 7.99 | 7.99 | 7.99 | 7.99 | 7.99 |
| G/T (dB/°K) | 11.32 | 11.32 | 11.32 | 11.32 | 11.32 | 11.32 |
| **LINK MARGIN TABULATION** | | | | | | |
| TERMINAL EIRP (dBW) | 108.10 | 101.60 | 95.08 | 76.06 | 71.62 | 65.80 |
| PROPAGATION LOSS (dB) | 231.85 | 231.85 | 231.85 | 231.85 | 231.85 | 231.85 |
| SATELLITE G/T (dB/°K) | 11.32 | 11.32 | 11.32 | 11.32 | 11.32 | 11.32 |
| BOLTSMAN'S CONSTANT (dB-Hz) | 228.60 | 228.60 | 228.60 | 228.60 | 228.60 | 228.60 |
| $C/N_o$ (dB-Hz) | 116.17 | 109.67 | 103.15 | 84.13 | 79.69 | 73.87 |
| REQUIRED $E_b/N_o$ (dB) | 10.00 | 10.00 | 10.00 | 10.00 | 10.00 | 10.00 |
| REQUIRED DATA RATE (Bps IN dB) | 80.00 | 80.00 | 70.00 | 60.00 | 50.00 | 50.00 |
| LINK MARGIN (dB) | 26.17 | 19.67 | 23.15 | 14.13 | 19.69 | 13.07 |

511140.0

### 1.2.2 Nulling Resolution

Intelligent choice of the amplitude and phase of the quiescent radiation pattern of an adaptive antenna has been shown to reduce the minimum jammer-to-user separation by more than a factor of three. This is illustrated in Figure 1-6 where the adapted and quiescent patterns, of a 5-beam antenna, are shown for two conditions of the quiescent pattern.

1. All beams are in phase.
2. Adjacent beams have a relative phase of 90 degrees.

The beams have a one-half power beam width of 0.5 degrees (i.e., an aperture diameter = 140 wavelengths) and they are spaced so they crossover at a point 6 dB below their peak. Notice at the -10-dB level, the width of the null decreases from about 0.6 degrees to 0.2 degrees.

This improved performance with a 140-wavelength MBA aperture is very close to the best that can be achieved with a TPA spanning a 450-wavelength aperture. Unfortunately, the MBA's control system must sense the general direction of a line connecting the user and the jammer and install the phase gradient in that direction in order to achieve results as good as those shown in Figure 1-6. Nulling resolution can be improved still further by amplitude tapering the quiescent pattern (see Figure 1-7); however, this latter improvement is at the expense of about 5-dB reduction in G/T over a large portion of the coverage area. This information may be readily available, or an algorithm could be devised to search for the best-phase taper by trying a limited number of tapers and selecting the best one. The TPA can achieve this same resolution without knowledge of the jammer's general location with respect to a user.

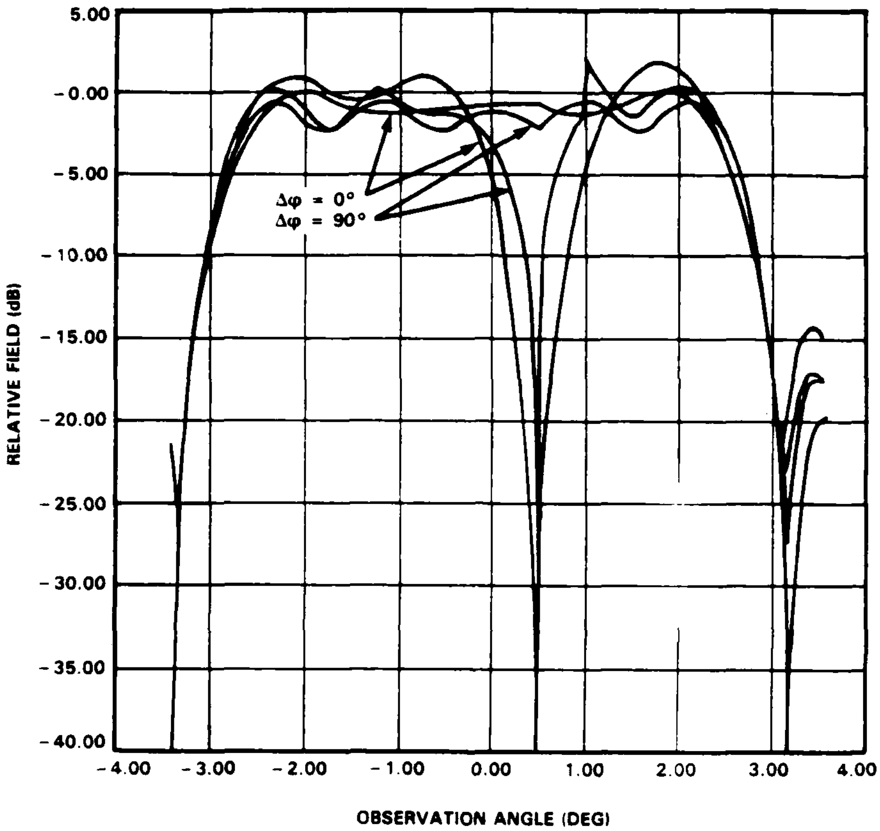



**Figure 1-6. Adapted Patterns: − 6 dB Crossover with $\Delta\varphi$ = 0°, 90°, D = 140λ**

511005.0

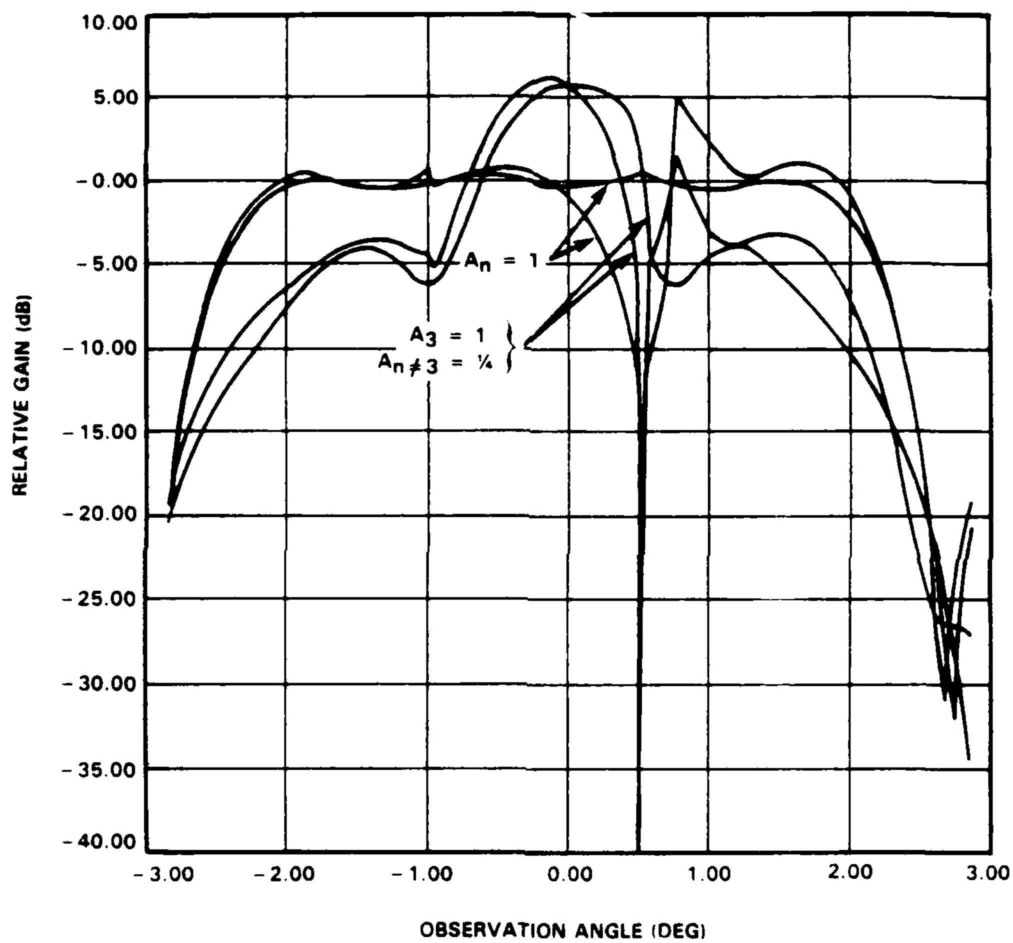



**Figure 1-7. Adapted Pattern*: $-6$ dB XOVER, $\Delta\varphi = 90°$, $D = 140\lambda$**

***$A_n$ = 1 : Phase Tapering Only.**
**$A_{n \neq 3}$ = ¼ : Phase & Amplitude Tapering.**

511006.0

### 1.2.3 Estimated Weight and Power

Tables 1-2 and 1-3 present estimated weight and power requirements for both SHF and EHF antenna systems. The dual-polarization SHF system payload weighs more than 300 lbs and requires more than 300 watts of power. The EHF package weighs 110 lbs and requires 112 watts of power. In view of this it is difficult to justify a dual-polarized system instead of an EHF package. The EHF package weight will increase to about 300 lbs and 300 watts when a processor and downlink antenna are added.

## 1.3 RECOMMENDATIONS

This study provides a first-order comparison of the relative merit of an SHF dual-polarized antenna versus an EHF antenna. The study indicates the salient areas for concern and tends to favor an EHF package over the dual-polarized SHF antenna. It indicates both are viable upgrades to DSCS III and, given the removal of NSSK fuel and addition of an on-board autonomous nulling algorithm, the present launch capability of DSCS III can support either package. In order to make a more sound judgment as to the detail configuration and performance requirements of either package, the following is recommended.

1. Design a strawman EHF antenna in sufficient detail to obtain more realistic performance, weight, and power estimates.*

---

*Work underway at Lincoln Laboratory on an EHF earth coverage and an area coverage MBA was briefed to the DSCS EHF Package Working Group Report [Reference 11].

**Table 1-2. Dual-Polarized Channels Estimated Weight and Power**

| QTY | ITEM | WT (lbs) | POWER (WATTS) |
|---|---|---|---|
| 1 | PHASE SHIFTER NETWORK | 50 | 5 |
| 1 | BEAM FORMING NETWORK (Rcv) | 75 | 5 |
| 1 | DEMUX | 5 | – |
| 3 | CHANNEL TRANSPONDER | 45 | 15 |
| 2 | 40 WATT TRANSMITTER | 40 | 240 |
| 2 | BEAM FORMING NETWORK (TRANS) | 40 | 5 |
| 1 | 10 WATT TRANSMITTER | 15 | 30 |
| – | WAVEGUIDE ETC. | 20 | – |
| 1 | JAMMER LOCATOR EQUIPMENT | 10 | 10 |
| | | 300 | 320 |

511141.0

## Table 1-3. Weight Budget : 285 Beam MBA

| ITEM | UNIT WT (lb) | POWER (W) | NO. REQ. | WT. (lbs) | POWER (W) |
|---|---|---|---|---|---|
| SWITCH | .0062 | .55 | 269 | 1.67 | 21.00 |
| WAVEGUIDE | .002 | | 7600 | 15.20 | .00 |
| WAVEGUIDE FLANGE | .002 | | 1200 | 2.40 | .00 |
| #4-40 × 3/16 SCREW | .001 | | 4460 | 4.46 | .00 |
| #4 LOCKWASH | .003 | | 4460 | 13.38 | .00 |
| FEEDHORN | .0008 | | 285 | .23 | .00 |
| POLARIZER | .01 | | 285 | 2.85 | .00 |
| SWITCH DRIVER & WIRING | .03 | | 269 | 8.07 | .00 |
| ANTENNA-LENS | 30 | | 1 | 30.00 | .00 |
| MISC. | | | | .00 | .00 |
| TOTAL ANTENNA AND SWITCH | | | 18829 | 78.26 | 21.00 |
| 16 DEGREE OF FREEDOM PROCESSOR | | | | 30.00 | 70.00 |
| CANDIDATE ANTENNA TOTALS | | | | 108.26 | 91.00 |
| ADDITIONAL SW. ISOLATOR | .0062 | | 285 | 1.77 | 21.00 |
| TOTALS | | | 19114 | 110.02 | 112.00 |

511142.0

2. Include in the strawman a detailed design of the antenna, switch system, nulling algorithm, signal processor, transmitting subsystem, control system, and necessary redundancy.

3. Consider the hybrid antenna configuration proposed in Section 3.2.3 in more detail.

4. Evaluate the polarization purity of the current DSCS III production antennas and the DSCS terminal antennas.

5. Analyze traffic effect on the potential of beam conflict in the switched-beam EHF MBA and probable self jamming in the dual-polarized SHF system.

6. Ascertain the range of G/T and difference in terminal EIRP available in response to the traffic model studied under point 5.

# CHAPTER 2

## ANTENNA DESIGN RECOMMENDATIONS FOR A DUAL-POLARIZED DSCS III SHF ANTENNA SYSTEM

Operational success of the DSCS III and development of the related terminal community has resulted in a need for an increase in the payload's communication capacity. Currently most up and downlinks are configured to provide FDMA; the EIRP of the multiple downlinks and the G/T of the multiple uplinks are large enough to accommodate increased communication capacity if associated terminals have sufficient EIRP and G/T. However, DSCS III will be bandwidth limited if the terminals have sufficient EIRP on their uplink and G/T on their downlink. Since the total bandwidth of the DSCS III payload is 375 MHz*, it cannot have a communications capacity greater than about 375 Mbps given rate one-half coded quaternary modulation (e.g., QPSK). The need to assign channels to communities that use FDMA can result in some channels being under utilized; this may, in a practical sense, reduce the maximum capacity to about 200 Mbps. Significant increases in terminal or payload EIRP or G/T in conjunction with higher-order modulation formats and higher code rates could be used to increase the throughput communication capacity. However, both the antijam and unprotected communication capacity may be doubled through use of a dual-polarized antenna system. This is because of the effective increase in bandwidth through frequency reuse possible with independent transmission on each of two orthogonally polarized radio waves. This technique (i.e., a dual-polarized antenna) should be considered as a primary technique for a DSCS III SHF upgrade, in addition to other improvement techniques. This study report addresses the design and salient performance characteristics of a dual-polarized antenna system.

*For DSCS III models up to B-7; for models B-8 and beyond, the available bandwidth increases to 405 MHz.

First a brief review of the pertinent characteristics of the DSCS III payload is presented. This is followed by a general description of the modifications and additions required to produce a dual-polarized system. A strawman dual-polarized system is then detailed and an estimate of its performance characteristics is presented. The latter includes self-jamming, jammer suppression through pattern nulling, and acquisition and synchronization considerations. These are followed by consideration of terminal modifications, etc., and propagation effects. Finally uplink power control and an estimate of the weight and power required to implement a dual-polarized system are addressed.

## 2.1 CURRENT ANTENNA SYSTEM

In this section the DSCS III receiving and transmitting antenna subsystems are described. Pertinent performance characteristics are discussed to establish a basis to which the dual-polarized system can be compared. In accordance with convention the receiving subsystem is discussed first followed by the transmitting subsystem.

### 2.1.1 Receiving Antenna

A 61-beam MBA and two earth coverage horns comprise the payload's uplink communication antenna suite. The earth coverage horns are connected directly to channel numbers 5 and 6 transponders (see Figure 1-1). Each of the 61 feeds of the MBA are connected to a beam-forming network (BFN) through a 360-degree phase shifter. The BFN consists of waveguide, coaxial transmission line, and 60 variable power dividers (VPDs). The output of the BFN is frequency division multiplexed into channel numbers 1, 2, 3, and 4 transponders.

The feed horns (Figure 2-1) of the MBA are RHCP and have a single output port. Each feed horn has a second port,

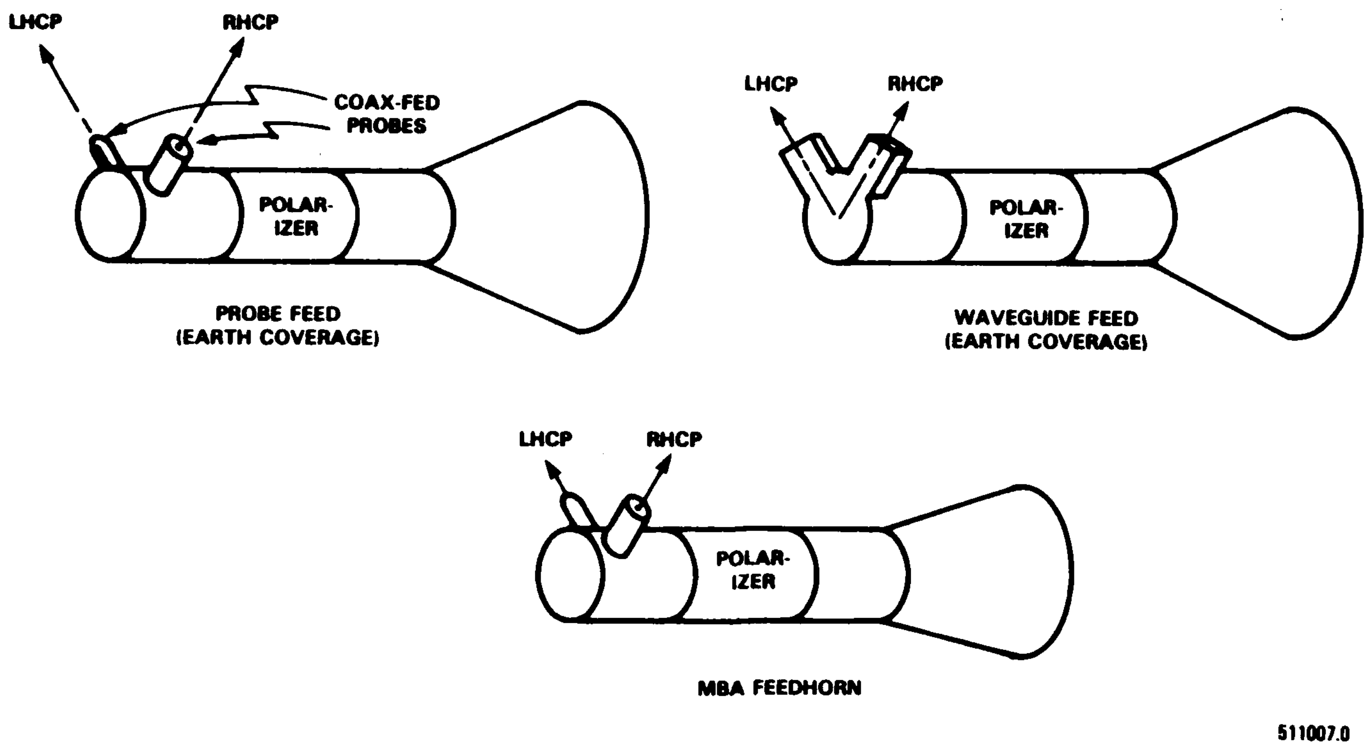



Figure 2-1. Antenna Horns

terminated in a matched load and coupled to a LHCP signal to improve the polarizer's performance. The phase shifters, BFN, and multiplexer are "single polarization" devices. That is, they transmit a single signal at a given frequency or over a given frequency band.

### 2.1.1.1. Description and Block Diagram

A simplified block diagram of the DSCS III payload is shown in Figure 1-1. The blocks labelled with channel numbers receive signals over the frequency band indicated, translate them to a corresponding downlink frequency band in the range 7250 to 7750 MHz, and amplify them. It is the bandwidth of these transponders that limits the current system's communication capacity. The BFN (see Figure 2-2) and phase shifters are configured through ground control. Knowledge of known user locations is utilized to establish a jammer-free or quiescent radiation pattern. The JLE uses the angular resolution property of the 61-beam MBA to locate interfering signal sources. In fact the JLE, in conjunction with ground equipment, is capable of locating all large signal sources in the FOV of the MBA. The JLE, through a solid-state switch and an array of 61 couplers, samples the total power received at each feed horn and hence from each beam coverage area. The amplitude of these power samples is digitized and telemetered to a ground control station. After computing the desired weights, or equivalently the phase shifter and VPD settings, the appropriate commands are telemetered to the payload. The weights are installed and the radiation pattern results.

### 2.1.1.2 Pertinent Performance Characteristics

The MBA lens is designed to operate with RHCP signals. However, it will operate equally well with LHCP signals (i.e., with crosspolarized signals) but requires two ports on each

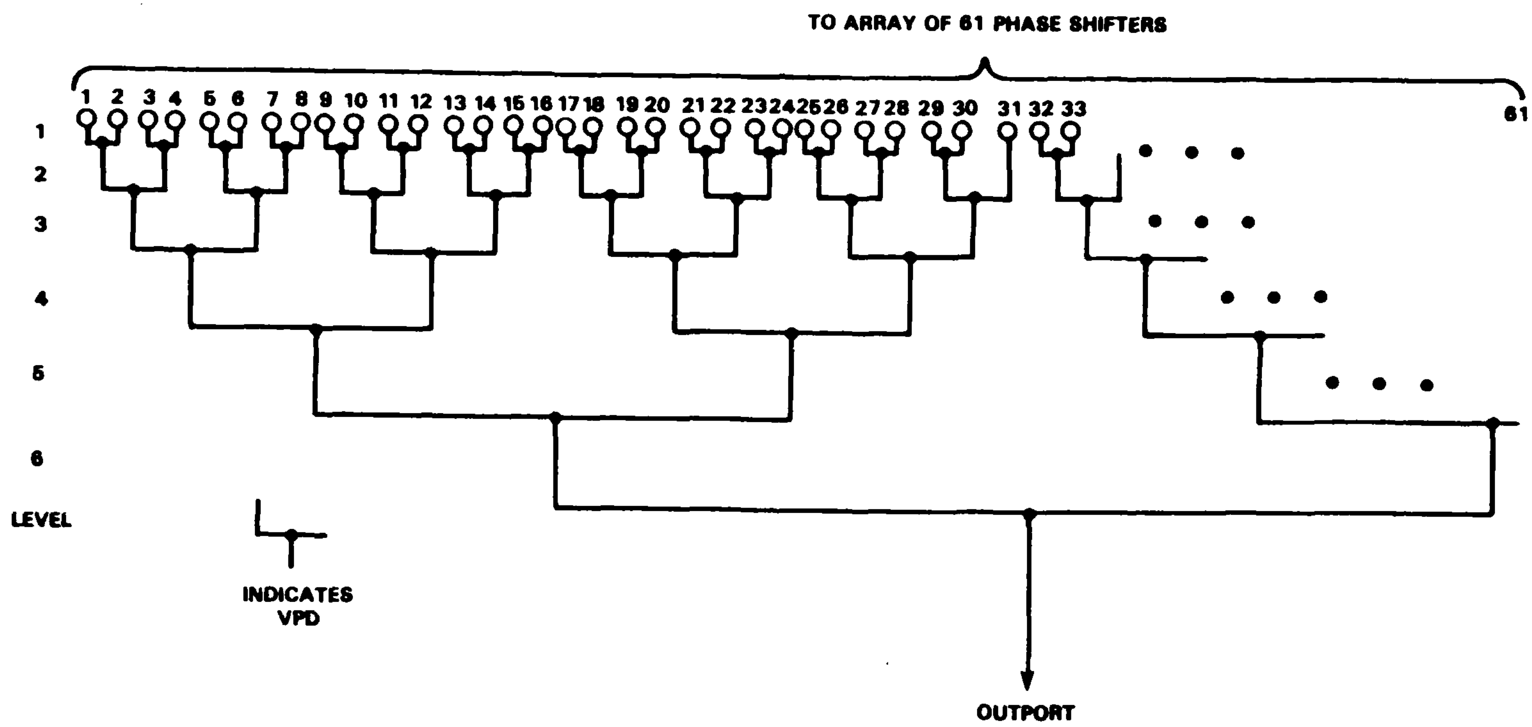



Figure 2-2. 61-Port Beam Forming Network

511008.0

feed horn: one for each polarization. The current design has these two ports.

The lens and the feed horns currently provide satisfactory performance over the frequency band 7975-8315 MHz. It is likely that performance of this same lens and feed horn array will perform adequately over the entire 7900- to 8400-MHz band if it were desirable to do so. It is also likely the phase shifters and the BFN operate satisfactorily over the entire 500-MHz frequency band. It is important to note the current JLE is sensitive only to RHCP signals.

#### 2.1.1.3 Beam-Forming and Nulling Algorithm

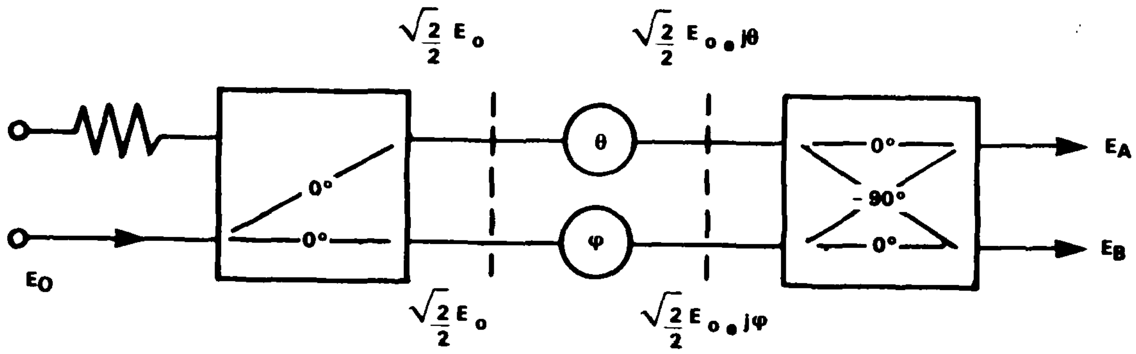
Although there are many different algorithms for determining the antenna weights, they all derive from a basic knowledge of the desired signal's structure (i.e., waveform), and the assumption that the jammer signal power is larger than the individual user's radiated power or larger than the total instantaneous power radiated by a group of users. The latter general algorithm is called the "power inversion" or Appelbaum-Howells algorithm. It requires the use of a spread-spectrum modulation to force a jammer to radiate signals over a frequency band, W, much larger than the user's information or instantaneous frequency band, R. The ratio W/R divided by the required energy per bit-to-noise power density ratio ($E_b/N_o$) is called the antijam processing gain and is approximately equal to the ratio of jammer-to-user power required to disrupt communication.

When pseudonoise or a (pseudo) random sequence is used to modulate the communication signal, the random sequence represents a known part of the modulated wave. The communication signals represent the unknown part of the modulated wave. The random sequence modulation is a much wider bandwidth signal than the information band; consequently, partial demodulation

of the received waveform (i.e., stripping off the known pseudonoise modulation) collapses the user's bandwidth while spreading the jammer's bandwidth. This increases the user's signal-to-jammer ratio significantly permitting separation of the user signal from the received waveform. The resulting "user free" signal can be used to identify jammer signals and shape the antenna pattern to reduce or remove them. (Such a technique requires on-board processing techniques beyond the scope of this report.)

The DSCS III nulling algorithm uses the known user locations to determine the desired quiescent pattern. Data received from the JLE is used to determine jammer location(s). The known antenna radiation pattern (actually "receiving" pattern), both amplitude and phase, are used to compute the desired weights using a power-inversion algorithm. When these weights are installed, the radiation pattern will have nulls in the direction of all jammers. Gain in the direction of the users will be dependent on the angular separation between a user and a jammer. For jammer-to-user angular separation greater than about 1.5 degree (approximately 60 percent of the 3-dB beamwidth), placing a null in the direction of a jammer will decrease gain to the user less than 3 dB.

Using calculated weights and known antenna radiation patterns, the ground control facility can calculate the expected radiation pattern. This permits alteration of the desired quiescent pattern to enhance coverage provided to user terminals in close proximity to jammers. This enhancement is limited to the fundamental resolving power of the lens; that is, the ratio of the diameter of the lens to the operating wavelength determines the minimum tolerable jammer-user separation. It is important to note the DSCS III JLE does not measure LHCP signals because they are rejected from the current feed horns. The VPDs (Figure 2-3) in the beam-forming network use latching ferrite phase shifters as a power-division control




$$E_A = \frac{E_O}{2}\left[e^{j\theta} + e^{j(\varphi + 90°)}\right] = E_O \sin\left(\frac{\varphi - \theta}{2} + \frac{\pi}{4}\right) \exp\left[j\left(\frac{\varphi + \theta}{2} - \frac{\pi}{4}\right)\right]$$

$$E_B = \frac{E_O}{2}\left[e^{j\varphi} + e^{j(\theta + 90°)}\right] = E_O \cos\left(\frac{\varphi - \theta}{2} + \frac{\pi}{4}\right) \exp\left[j\left(\frac{\varphi + \theta}{2} - \frac{\pi}{4}\right)\right]$$

**Figure 2-3. Variable Power Divider (VPD)**

511009.0

element. In order to obtain the desired accuracy of the insertion phase, any change in power division is a two-part process. Each phase shifter is first driven into saturation with a "set" pulse. In this "set" state the insertion phase of a VPD equals 45 degrees relative to the operational state, and the input power divides equally between the output ports. Next the "reset" pulse is applied, the desired power division is obtained, and the insertion phase of the VPD reduced to zero (i.e., the operational state). In order to prevent this transient "set pulse" state from introducing phase error in the communication signal and intolerably reducing power received from a user, the 60 VPDs are switched in accordance with an algorithm invented by Dr. Alan Simmons at Lincoln Laboratory. This algorithm has two major steps:

1. Beginning with the first level of VPDs shown in Figure 2-2 (i.e., the VPD closest to the feed horns) each variable power divider is set to its new value if the G/T provided a user in both the pre- and post-configuration change states is either an increase in G/T or is no more than 3 dB lower than the G/T that will be provided in the new state. The VPDs in the first level are set to provide equal-power split if the foregoing condition cannot be satisfied.

2. Next, the VPDs in the remaining levels are set in accordance with the same rule. Then beginning with the highest level (i.e., the one closest to the BFN output port) of VPDs and proceeding toward the first level the VPDs are set to their final or new value of power division.

In short the algorithm establishes an intermediate pattern which reduces, by less than 3 dB, the G/T provided to users that remain in the system through the configuration change.

The phase shifters are all switched at the same time; consequently during the set/reset pulse period changing the phase shifters may introduce an appreciable change in the characteristics of a prescribed null in the radiation pattern; it should not have a significant effect on the quiescent pattern. This transient condition lasts for less than 6 microseconds per level in the VPD tree and will probably permit jammer signals to interfere during this period. This undesirable effect can be minimized by interleaving bits over a 6-microsecond span. At a data rate of, say 10 Mbps, this requires a 60-bit storage device: a small increase in hardware in the terminals. However, most DSCS spread-spectrum data rates have data symbol durations which are much greater than 10 microseconds and therefore would be unaffected by the switching of a single level of VPD.

A change in the insertion phase of the communication signal can be precomputed and compensated for by means of a single-phase shifter at the output of the BFN. In fact, a two-bit phase shifter (i.e., zero degrees and -45 degrees) located between the BFN and the multiplexer, functions to compensate the 45-degree insertion phase introduced while switching a level of VPDs.

Thus the beam-forming algorithm consists of two passes through the seven levels of VPDs in the BFN, one level at a time. While switching each and all levels, a compensating phase shift is introduced at the output of the BFN to prevent phase modulation of the communication signals during reconfiguration of the MBA's radiation pattern. Transient effect of switching the phase shifters has essentially no significant effect on the quiescent patterns but does introduce a transient change in jammer suppression.

#### 2.1.1.4 Reconfiguration Time

If the desired weights are known, the antenna can be reconfigured in less than 30 seconds. Most of this reconfiguration time is required to transmit the commands to the payload through the associated decryption devices. Once the payload receives the set of commands, the BFN and phase shifters can be set to their new values in less than a few hundred microseconds.

Determination of the MBA pattern-forming weights can require up to several minutes depending on the complexity of the performance characteristics. If nulling is not required, calculation of the weights can be completed in a few minutes. This time approximately doubles when a null must be installed. Additional time is required to compute the desired weights if jammer location must first be obtained using data collected by the JLE. First the JLE must receive signals exceeding a preset threshold and then begin sending appropriate data to the control facility. Using known MBA radiation pattern characteristics, the power received in each beam of the MBA and an appropriate algorithm, the location and relative strength of jamming source is determined. This process can take up to approximately 5 minutes depending on the relative strength and location of the jammers. For some worst-case scenarios, up to approximately 10 minutes can pass between the time a jammer is first suspected to be present and before its signals are suppressed by installation of a null in the radiation pattern.

### 2.1.2 Transmitting Antenna

The downlink antenna suite consists of two 19-beam MBAs, a GDA and two earth coverage horns. Several redundancy switches permit connection of channels number 1 and 3 to one 19-beam MBA, channels number 2 and 4 to the other 19-beam MBA, channels numbered 1, 2, or 4 to the GDA, channels numbered 3 or 5 to one

of the earth coverage horns, and 4 or 6 to the second earth coverage horn.

### 2.1.2.1 Description and Block Diagram

Referring to Figure 1-1, both MBAs are connected to 40-watt TWTAs (primary connection) via transfer switches and a 1:19 BFN. Each BFN is similar to that used on the uplink except only 18 VPDs are required for each BFN and, with the exception of those used in the VPDs, phase shifters are not required.

A GDA is used to provide a higher EIRP downlink (i.e., higher than with the MBA). Two earth coverage horns provide wide area coverage but at decreased EIRP.

### 2.1.2.2 Pertinent Performance Characteristics

Reconfiguration of the radiation pattern of each MBA is carried out using the Lincoln Laboratory beam-forming algorithm described in Section 2.1.1.3. However, the time required to complete a reconfiguration is less than a minute; principally because there is no need for a pattern with a prescribed null. Although the lens, feeds, and BFN could operate over the entire 7900- to 8400-frequency band, they were designed for operation over the 7975- to 8315-MHz band. Operation outside their design frequency band will result in some degradation in performance but perhaps this degradation is tolerable.

Switches connecting the input port to either of two output ports are indicated by a circle enclosing the letter "S." A square enclosing a "D" is used to represent a dual-frequency multiplexer. A circle enclosing a "T" represents a three-frequency multiplexer or more accurately a band-stop filter or trap.

The GDA provides a 3-degree diameter coverage area with EIRP = 44 dBW in channels 1 or 2, and 37.5 dBW in channel 4. It can place the 3-degree coverage anywhere on the earth disk. It is included in the antenna suite to alleviate the need to split an MBA radiation pattern (and suffer a concomitant reduction in EIRP) in order to provide service to two separate communities unlikely to be colocated.

The earth coverage horn antennas are similar to those on the uplink. These and all downlink antennas radiate LHCP signals. These downlink antennas can transmit either sense of CP with a straight forward change in the excitation port of the feed horns of the MBAs. The MBA feed horns have a second port (currently terminated in a matched load) that can receive or transmit RHCP signals. The GDA feed can be modified to include ports for RHCP and LHCP signals. The earth coverage horns can be dual polarized by adding a LHCP excitation port.

#### 2.1.2.3 Beam-Forming Algorithm

The BFN distributes transmitter output signals to the feed horns to produce the desired pattern. Since there is no advantage to placing a null in the transmit radiation pattern, adjusting the phase of the signals exciting each feed horn of the MBAs provides negligible, if any, advantage in the antenna system's performance.

In order to prevent unnecessary phase and amplitude transients in the communication system, the Lincoln Laboratory procedure for installing weights is used to reconfigure the antenna. It is identical to that described in Section 2.1.1.3.

#### 2.1.2.4. Reconfiguration Time

Without the need to determine jammer location (for downlink transmission) and the concomitant weights for a radiation

pattern with an appropriate null, the desired BFN weights can be calculated and installed in less than one minute. Clearly the time to reconfigure the uplink MBA dominates the reconfiguration time.

## 2.2 DUAL-POLARIZED SATELLITE SHF UPGRADE ANTENNA SYSTEM

The DSCS III antenna can be made capable of radiating and receiving both RHCP and LHCP signals by connecting the appropriate transmission line to already existing antenna ports. Choosing the correct transponder configuration, channelization, high-power amplifier and beam-forming or nulling algorithm poses the widest range of choices and the most challenging design problem. This section addresses the antenna design, presents some transponder configurations with supporting rationale, and discusses the incorporation of adaptive antenna-pattern suppression of interfering signals.

In principle, acquisition of and synchronizing with received signals can be as with the current system. Conditions necessary for this conjecture will be discussed.

The current system carries a substantial amount of fuel to maintain the satellite's axis coincident with the earth-satellite axis (i.e., NSSK). Frequent reconfiguration of the MBAs and pointing of the GDA could substantially reduce the fuel required for NSSK. This savings in weight might be used for other purposes such as increased transponder capacity or installation of an additional communication package operating at EHF. Potential reduction in weight will be discussed in this section.

### 2.2.1 Antenna Design

The DSCS III MBAs each have dual-polarized feed horns (Figure 2-1) in that they are circularly polarized and have a

port that couples to orthogonally polarized signals. This port is terminated in a matched load to improve the circularity of polarization; replacing the matched load with a matched receiver or an isolator (with a matched load) will permit operation as though a matched load were attached to the port. Hence connecting a receiver or a transmitter to these orthogonal ports converts the receiving and/or transmitting antennas into dual-polarized devices. It is important to note only the antennas can be dual polarized; the phase shifters, BFNs, frequency demultiplexers (DMUXs), multiplexers (MUXs), switches, and filters are all "single" polarization devices. This is because they can only carry one signal at any instant of time, over a given frequency band.

The GDA can be made dual polarized by changing its feed horn. This is straightforward and usually consists of adding a probe feed, or waveguide feed, at the excitation end of the feed horn. It is common practice to include the orthogonal port (e.g., the waveguide feed or probe feed) in the design of a circularly polarized feed and terminate it in a matched load. Since, the polarizer, the feed horn aperture, and the reflector are fundamentally dual polarized; CP signals with opposite polarization can be simultaneously received (or transmitted) at the orthogonal port.

Earth coverage horns are essentially the same type of device as a feed horn for the MBA or the GDA. They are fundamentally dual polarized; they will be dual polarized if one port for each polarization is provided. As with the other DSCS III antennas, the polarizer converts linearly polarized signals into circularly polarized signals. Furthermore orthogonally polarized signals, either linear or circular, exist on a one-to-one basis. That is, if vertically polarized signals convert into RHCP signals then horizontally polarized signals convert into LHCP signals. Consequently the addition a of linearly polarized exciter will convert the earth coverage

horns to dual-polarized devices. This general characteristic of circularly polarized horns is demonstrated in Figure 2-1 where both coax-probe-fed and waveguide-fed horns are shown. For both configurations the polarizer and horn are essentially the same. Note that each exciter is sensitive to primarily one of two orthogonal linearly polarized signals. The polarizer converts that linearly polarized signal into either RHCP or LHCP.

In summary, any of the DSCS III communications antennas can be converted to dual-polarized devices in a straightforward manner. In some instances it may be as simple as replacing a load with a receiver or a transmitter. The most complicated conversion would require the addition of an "orthogonal" port to the current design and making this a straightforward use of state-of-the-art technology.

### 2.2.2 Transponder Design

Upgrading DSCS III to have dual-polarized antennas will increase its throughput communication capacity by virtue of the increase in bandwidth realized through frequency reuse. Just how this increased "bandwidth" is to be allocated to the current and potential users of the DSCS III SATCOM system is subject to future Joint Chiefs of Staff (JCS) policy decisions. Nevertheless one can postulate that use of this increased bandwidth need not require a frequency plan and channel allocation identical to that currently used on DSCS III. In fact it may prove wise to move only the higher-data-rate users to the orthogonally polarized channels since that might lead to minimum impact on the terminal community as a whole and this change in frequency assignment would free substantial bandwidth currently used by the high-data-rate users. The latter are potentially the only users that can efficiently use a single, or a few, wideband channels.

Conversely the smaller, low-data-rate users have a greater need for improved EIRP or G/T; bandwidth currently available to them may be inadequate principally because of multiple access constraints. A study of the current and predicted communication traffic associated with the various communities is beyond the scope of this study but would be very helpful in assigning channel characteristics to the transponders attached to the crosspolarization ports of the antennas.

As an example of the possible variations on transponder design, assume that increased bandwidth (i.e., the added cross-polarization channels) would be assigned to the high-data-rate users and they would in turn give up channels they use in the current frequency plan. Therefore the high-data-rate users could have a total bandwidth of 340 MHz, two 155-MHz channels, a duplication of the current channels 1, 2, 3, and 4, or any mixture that gives ample guard bands and has a total bandwidth equal to that currently occupied by channels 1, 2, 3, and 4.

The total bandwidth associated with the added polarization may be limited to 340 MHz because the uplink MBA is designed to operate over the band 7975-8315 MHz. It can probably operate satisfactorily over the entire 7900- to 8400-MHz band; that is the entire 500-MHz frequency band with slight degradation in performance at the edges of the band. The current phase shifters and the BFNs can probably operate over the entire 500-MHz band with slight degradation in performance. The DMUX and MUX filter designs could either be used directly or appropriately modified. These devices and the switches could meet the requirements of either a 340- or 500-MHz band; their design and design procedure are well known. On the other hand, the downlink MBAs are designed to operate over a 255-MHz band and would probably have serious degradation of performance if they are operated over a 500-MHz band. It is likely, however, that either one of the current 19-beam MBAs could operate satisfactorily over a 340-MHz band. These units have narrower

operating bandwidth than the uplink MBA because of more stringent gain requirements. In view of the foregoing, the strawman configuration shown in Figure 1-2 will be chosen for study. Two channels are chosen so that TWTAs and filters in the current design can be used with little, if any, modification. This allows the wideband users at least two point-to-point, or community-to-community, scenarios in place of the current system's capabilities. That is the wideband users would use either, or both, of the downlink MBAs. Each of the channels would have a bandwidth of about 155 MHz and a 40-watt high-power amplifier.

In order to set some bounds on individual user bandwidth, etc., consider a strawman earth satellite link consisting of an earth terminal using a 30-foot diameter paraboloid antenna, a one-kilowatt high-power amplifier, and a 500-degree Kelvin ($^{o}K$) receiver noise temperature. The satellite is assumed to have an uplink antenna gain greater than 25 dB; with a system noise temperature of 1000 degrees, a 40-watt downlink TWTA output and downlink antenna gain equal to 27 dBi. That is, each of two downlink MBAs is configured to produce a single pencil beam in the user(s) direction. The uplink MBA has a pattern shaped to accommodate all crosspolarization users; consequently its gain will be less than the maximum possible (i.e., 32 dBi). In view of the large uplink signal-to-noise ratio, it seems reasonable to include a 5-dB degradation in uplink gain to allow for pattern shaping. The resulting up and downlink budgets are given in Table 2-1. Notice that up to four 40-Mbps users,(i.e., 160-Mbps total) with a 10-dB signal-to-noise (S/N), can be accommodated on each of the downlink MBAs and a total of eight 40-Mbps users can be supported on the single uplink MBA. Addition of this crosspolarization configuration will permit about 300 Mbps (depending on modulation and required $E_b/N_o$) increase in throughput capacity. The large (27-dB) uplink margin could be reduced to approximately 10 dB decreasing the uplink TWTA output from a kilowatt to about 20

**Table 2-1. Link Budget for Strawman Crosspolarization Channels**

| | CHANNELS 7, 8 | | CHANNEL 9 ** | |
|---|---|---|---|---|
| | UPLINK | DOWNLINK | UPLINK | DOWNLINK |
| ANTENNA GAIN (dB) | 57 | 27 | 43 | 17 |
| CIRCUIT LOSSES (dB) | −2 | −2 | −2 | −1 |
| TRANSMITTER (dBW) | 30 | 16*** | 23 | 10 |
| EIRP (dBW) | 85 | 41 | 67 | 26 |
| $\lambda^2/4\pi R$ (R = 20,000 MILES) (dB) | −200 | −200 | −200 | −200 |
| ATMOS. ATT. (dB) | −3 | −3 | −3 | −3 |
| | −118 | −162 | −118 | −177 |
| ANTENNA GAIN (dB) | 32 | 57 | 15 | 43 |
| PATTERN SHAPING LOSS | −5* | −− | −− | −− |
| LOSSES | −2 | −2 | −1 | −2 |
| RECEIVED POWER (dB) | −83 | −107 | −104 | −138 |
| k (dB) | −226 | −226 | −226 | −226 |
| T (dB) | 30 | 27 | 30 | 27 |
| B (dB) | 76 | 82* | 63 | 53 |
| kTB = No | −120 | −123 | −133 | −127 |
| $S/N_o$ (dB) | 27 | 10 | 29 | 11 |
| $E_b/N_o$ (dB) | 10 | 10 | 10 | 10 |
| EXCESS LINK MARGIN | 17 | 0 | 19 | 1 |

511143.0

* EQUIVALENT OF FOUR TERMINALS @ 40 Mbps/LINK.
** SEE FIGURE 1-2
*** ASSUMING 3 dB BACKOFF FOR LINEARITY

watts. This would reduce the uplink antijam capability by 17 dB, and therefore is a peacetime operational configuration.

Alternatively the terminal could use a 1-kW TWTA to have the higher antijam capability and the ferrite phase shifter, and the BFN could be replaced with diode-phase shifters and a strip-line BFN. The use of diodes and strip line would decrease the weight substantially (about 30 lbs) and increase circuit losses from about 3 dB to about 10 dB. Increases in loss reduces S/N proportionately, has negligible effect on antijam capability and only low-risk development is required, since all components are readily available. Terminals with significantly less EIRP (i.e., less than 78 dBW) could only be supported if they had a proportionately lower data rate. In other words the proposed inefficient phase shift, BFN network, could only support terminals operating at 40 Mbps if they have greater than 15-foot diameter antennas and TWTAs with greater than 500-W average RF output power.

Using diodes and strip line in the downlink BFNs would increase the circuit losses and decrease the transmitted power. This causes a proportionate reduction in EIRP and a concomitant decrease in the maximum data rate that can be supported. In short, Table 2-1 indicates the crosspolarization payload is limited by its EIRP on the downlink; hence, ferrite phase shifters and low-loss transmission line (e.g., waveguide and/or coax) should be used instead of diode-phase shifters and a strip-transmission line.

If weight and power growth is permitted, as is being considered for EHF options, it is important to consider further increases in communication capacity by adding a transponder between the crosspolarized ports of the channel 6 earth coverage horn antennas. If a 10-watt TWTA is used, channel number 9 (see Figure 1-2) could support a 0.2-Mbps communication data rate to a terminal with six-foot diameter antenna aperture

and a 500-degree receiver noise temperature. (The data rate that could be supported increases as the square of the diameter of the antenna aperture and in proportion to the increase in power output of the TWTA.)

### 2.2.3 Self-Jamming Considerations

When two terminals access the satellite at the same time using the same frequency band with crosspolarized signals there is a potential for them to jam one another. The satellite's ability to separate the incident flux into copolarized and crosspolarized channels is the principal method of reducing, or eliminating, this form of self jamming. Just as filtering and timing prevent self jamming in FDMA or TDMA systems, purity of polarization is the key performance parameter in dual-polarized or frequency-reuse systems. Polarization purity or matching is obtained and maintained by appropriate design of the terminal and communication antennas and accommodation to the atmospheric, or propagation, effects on the electromagnetic waves that carry the signals between the two of them on an earth-satellite link. In this section a fundamental relationship, used to characterize self jamming, is developed. First, rationale for selecting polarization is reviewed.

When linear-polarized antennas are used by both the terminal and the satellite, each terminal must orient its direction of polarization to match that of the satellite. Although knowledge of the orientation and location of the satellite are sufficient to calculate the necessary polarization direction, terminals designed to accommodate, or match, their polarization to that of the satellite may be unnecessarily complicated. Usually system architects choose to operate using CP signals to eliminate this need to orient the direction of polarization. Furthermore opposite sense CP antenna exciters are orthogonal, that is they are uncoupled. On satellites or terminals that use the same antenna for

transmission and reception, it is common practice to transmit signals of one sense and receive signals of the opposite-sense CP. This provides about 25-dB isolation between transmit and receive signals and reduces the isolation required by the diplexing filters that must separate the transmit and receive signals from one another. DSCS III antennas receive RHCP on the uplink and transmit LHCP signals on the downlink for the foregoing reasons. The proposed dual-polarized system adds the capability for the spacecraft to receive LHCP signals on the uplink and transmit RHCP signals on the downlink.

Unfortunately, neither the terminal nor the satellite antennas transmit and/or receive CP waves. Rather, they are elliptically polarized; they are sensitive to both RHCP and LHCP waves where, for a "RHCP" antenna, gain for RHCP incident signals is about 25 dB greater than for LHCP incident signals. This ratio of gain for crosspolarized to copolarized signals (C) is directly related to the axial ratio of the antenna's polarization and will be discussed later. However, for frequency reuse using a dual-polarized antenna, C can be the principal mechanism that determines coupling between channels connected to RHCP and LHCP ports of the same antenna (e.g., as indicated in Figure 1-2).

Consider channels 1 and 7 (Figure 1-2), RHCP uplink signals will couple to both channels with those in channel 1 1/C stronger than those in channel 7. Similarly LHCP uplink signals will couple predominantly to channel 7 but they will also couple to channel 1 and channel 7 signals will be 1/C larger than channel 1 signals. Let P1(1-C) represent the power in the uplink LHCP wave of intensity P1, that is coupled to channel 1 and P7C represent the power in the uplink RHCP wave of intensity P7 that is coupled to channel 1. (P1, P7, and C must be expressed as numerics.) Assume these signals occupy the exact same bandwidth but are statistically independent.

Then considering only $N_oB$, the total noise power in the same bandwidth, and P7C as the only two undesirable signals in channel 1, the S/N ratio is given by,

$$S/N = P1*(1-C)/(P7*C+N_oB) \quad (1)$$

which can be written

$$S/N = (S/N_oB)*(1+(C/(1-C))*(P7/P1)*(S/N_oB))^{-1} \quad (1a)$$

Note that if the second term in the denominator of (1a) is small compared to 1, the self-jamming effect is negligible. However, if it is large compared to 1, the S/N ratio becomes,

$$S/N = ((1-C)/C)*P1/P7 \quad (2)$$

That is, C, the cross coupling, between channels 1 and 7 determines S/N and the system noise temperature. Using (1a) S/N was calculated as a function of P7/P1 for $S/N_o$ equal to 10, 20, and 30 dB. The results are shown in Figure 2-4.

If the uplink wave is elliptically polarized either because the terminal antenna has an axial ratio (axial ratio equals ratio of maximum-to-minimum response of a linearly polarized antenna when its orientation of polarization is rotated through 360 degrees) greater than 1 or because the propagation path depolarized the wave, the S/N will be degraded in accordance with

$$S/N = (S/N_o)*(1+((1-C')/C')*(P7/P1)*(S/N_o))^{-1} \quad (3)$$

where

$$C' = ((AR'-1)/(AR'+1))^2 \quad (4)$$

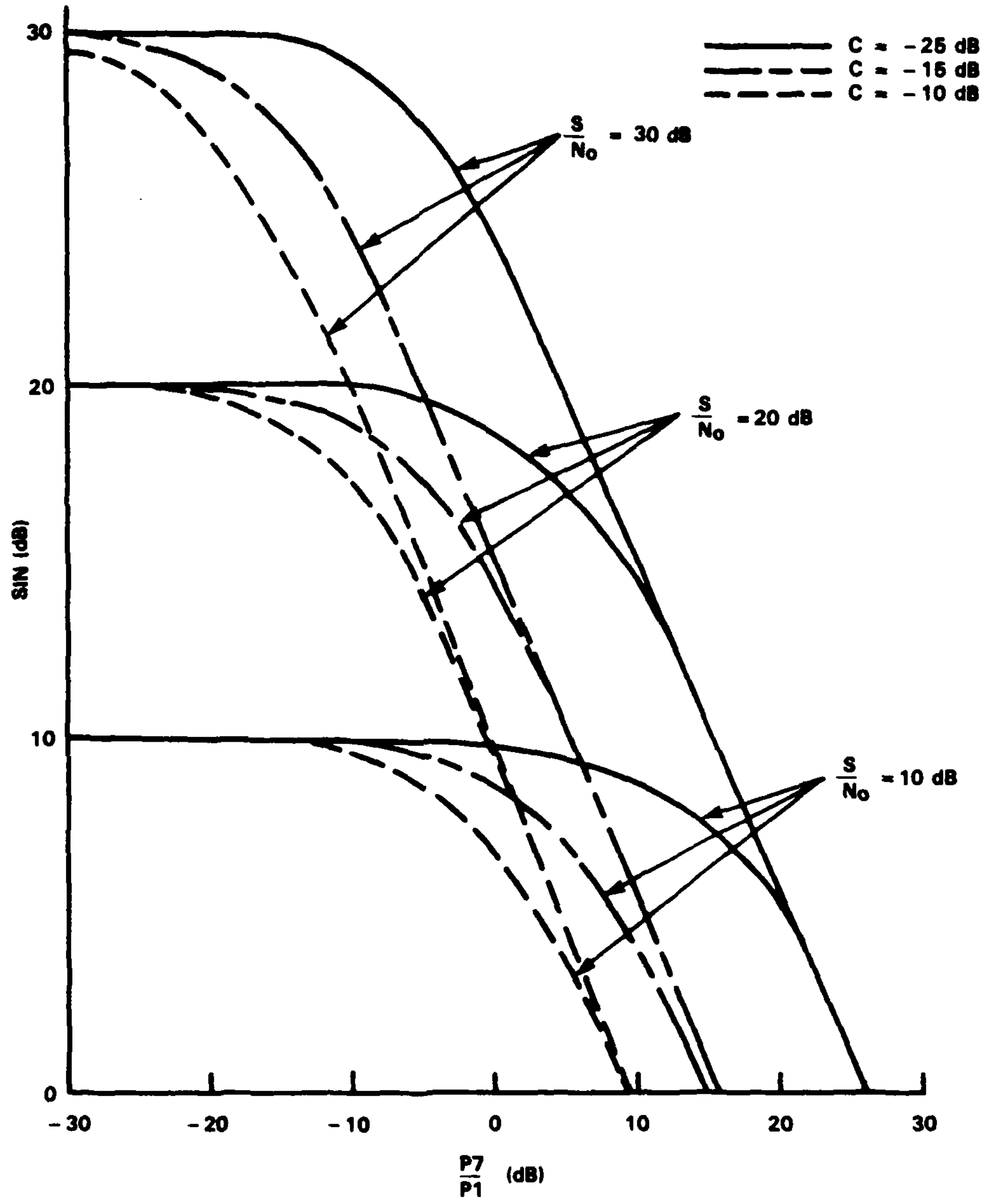



**Figure 2-4. Overall System Signal-To-Noise Ratio**

511010.0

and AR' is the axial ratio of the uplink wave. In order to sort out the three components of crosscoupling due to depolarization define AR, AR', and AR" as the axial ratio of the space- craft uplink antenna, the depolarization due to the propagation path, and the terminal's uplink antenna, respectively. The crosscoupling coefficients C, C', and C" are then computed using (4) and the corresponding value of AR, AR', and AR".

In order to obtain an end-to-end crosscoupling coefficient $C_{eff}$ it is necessary to combine all three effects. As long as C, C', and C" are small,

$$C_{eff} = C + C' + C'' \qquad (5)$$

The exact value of $C_{eff}$ can be calculated if the phase shift of the distorting medium is known. It will be approximately equal to that given by (5).

### 2.2.4 Performance Requirements

Continuing to address the strawman version of an SHF dual-polarized upgrade antenna system (Figure 1-2), it is important to establish some essential performance characteristics and a basis for them. In the foregoing section self jamming due to polarization impurity or deterioration was addressed. Self-jamming effects can also be caused by power robbing, imperfect filtering for FDMA, intermodulation noise introduced by non-linearity in the high-power amplifier and other devices, and multiple scatter effects that may alter polarization. The list can be even larger; however, current consideration will be limited to those devices, etc., that are peculiar to the addition of channels carrying crosspolarized signals.

The strawman approach proposes to use the same antenna for RHCP and LHCP signals. Consequently the output (input) ports

will provide good isolation (approximately 35 dB) between these channels. However, the terminal antenna will, in general, transmit (receive) signals polarized slightly different from those received (transmitted) by the payload. As discussed in the previous section, antennas on these two "terminals" and the propagation path between them introduce a mechanism for coupling "crosspolarized" signals into a channel. It was also shown that coupling via this mechanism can be expressed in the form:

$$C = ((AR-1)/(AR + 1))^2 \tag{6}$$

where AR is the axial ratio associated with either the satellite or ground terminal antennas and/or the polarization degradation introduced by the propagation path. Since DSCS terminals and satellite antennas have an axial ratio less than 1.25 (i.e., less than 2 dB) crosscoupling between channels, worst- case coupling between the antennas will be on the order of -13 dB. If the axial ratio of either the satellite or ground terminal antenna equals 1, crosscoupling, due to a 2-dB axial ratio of the other antenna, will be reduced to -19 dB. Cross- coupling due to polarization mismatch, when one terminal has an AR = 1, is given in Figure 1-4. When more than one source of crosscoupling can be characterized by a meas..ed, or estimated, axial ratio, overall crosscoupling (e.g., terminal and satellite antenna axial ratio greater than 1.0) can be estimated using:

$$C = 20\mathrm{Log}\left[\sum_{n=1}^{N} (AR_n-1)/(AR_n+1)\right] \tag{7}$$

Equation 7 shows that N sources of polarization degradation produce a worst-case coupling that is 6N dB greater than a single source.

Referring to published data and analysis, crosscoupling due to rain is about -15 dB for very heavy (approximately 20 mm/hr) rain rate (see Reference 5). Using equation (6), antennas with AR = 1.25 (2 dB) and heavy rain, under a most unfavorable combination, can result in -8-dB coupling between channels and the concomitant degradation in S/N given in Figure 2-4.

In the previous section, results shown in Figure 2-4 were determined using a simplified coupling model. Even minor power imbalance between uplink terminals can cause serious degradation of the channel used by the weaker terminal. For example two terminals with equal EIRP, $S/N_o$ = 10 dB and $C_{eff}$ = -10 dB could, in accordance with Figure 2-4, reduce each other's S/N about 3.5 dB. If one terminal has 10 dB more EIRP than the other, S/N of the weaker terminal will be degraded 10 dB. If $S/N_o$ of each terminal is 30 dB, a 10-dB imbalance in EIRPs will degrade the weaker terminal 30 dB. In other words, for $C_{eff}$ large (i.e., approximately -10 dB), S/N for the weaker terminal is reduced by more than the difference (in dB) of their EIRP.

In view of the foregoing and Figure 2-4, $C_{eff}$ should be less than -15 dB for barely tolerable performance and should be less than -25 dB if at all possible. For $C_{eff}$ less than -25 dB, terminals with $S/N_o$ = 10 dB will experience a 1.5-dB reduction in S/N if a terminal on the crosspolarized channel, operating over the same frequency band, has 10 dB greater EIRP. For the same scenario, except with $S/N_o$ = 20 dB, the weaker terminal will experience a 6-dB reduction in S/N. It is important to note that current single polarization operation requires disciplined EIRP control to prevent "power robbing," another form of self jamming. This same discipline could be applied when dual polarization is used; for example, $C_{eff}$ = -15 dB may be adequate if S/N greater than 9 dB is sufficient given S/N = 10 dB was desired.

When the conflicting terminals occupy different but overlapping frequency bands, the terminal operating over the larger band can, and should have a higher EIRP in accordance with:

$$EIRPa = EIRPb * Ba/Bb \tag{8}$$

where EIRPa, EIRPb, Ba, and Bb are the EIRP and frequency bandwidth of terminal a and terminal b respectively.

Clearly polarization purity (orthogonality) is the dominant factor in a dual-polarized system. Replicating circuits, amplifiers, filters, etc., is straightforward but maintaining the desired polarization purity can be very difficult. With any antenna, coupling between crosspolarized signals will be inherently less than -30 dB. However, matching the polarization of a terminal antenna to spacecraft antenna is difficult if crosscoupling must be less than -25 dB (i.e., axial ratio less than 1 dB). As discussed above, rain along the propagation path can depolarize the wave so as to degrade coupling between orthogonally polarized antennas to -15 dB. (See Reference 5). More will be said about this in Section 2.4.2.3.

Performance characteristics of all components and subsystems must be the same as for the current DSCS III payload. In other words, the crosspolarized channel components can be identical to, or modifications of, the current components except for a possible change in frequency bandwidths and frequency.

### 2.2.5 Pattern-Nulling Considerations

Currently DSCS III uses an array of phase shifters (Figure 1-1) and VPDs to weight the signals received on each of the 61 uplink beams of the MBA. These weighed signals are summed by the BFN (Figure 2-2) to produce a receiving pattern

with "nulls" in the direction of interfering signals while maintaining, as well as possible, a desired radiation pattern shape to accommodate user terminals. The current adaptive nulling system includes earth-based facilities to determine the location of interfering sources, calculate the desired weights, and command the uplink antenna weighting circuits accordingly. In this section this reconfiguration function will be considered for the crosspolarization channels and for both ground-control processing and on-board satellite processing.

### 2.2.5.1 Open Loop

The current method of adaptively changing the uplink 61-beam MBA's radiation pattern is conventionally referred to as an open-loop process. That is, the jammer locator measures the power received on each beam and transmits this data to a ground facility. Using known radiation patterns of the MBA (i.e., measured prior to launch) the jammer locator measured data, and the desired interference free (quiescent) radiation pattern, the ground facility calculates and commands the installation of an "optimum " set of weights. This set of weights is optimum in the sense the resultant radiation pattern is a best root mean square (RMS) error fit to the quiescent radiation pattern with nulls in the direction of the interfering sources. This is an open-loop process because there is no subsequent measurement of the performance of the installed pattern with continued correction to improve the measured performance. If all devices were error free and the calculation process free of approximations there would be no need for measurement of the resultant performance; DSCS III is reasonably free of these type errors and the resultant radiation pattern provides more than the desired interference suppression. However, the time delay between sensing the presence of an interfering signal and suppressing it can be undesirably long. Furthermore, the installed radiation pattern must have its nulls shaped so that spacecraft motion will not move the "null" off the direction to

the interfering source. Shaping the null decreases the depth of null possible and reduces the gain to users located close to an interfering source.

Referring to Figure 1-2, notice the array of phase shifters and the beam-forming network (i.e., VPDs) are identical for both the current DSCS III and the proposed cross-polarized channels. Consequently, the RHCP channels can install radiation pattern shapes completely independent of the radiation pattern installed by the LHCP channels. It follows that the LHCP channels could have a jammer-locater subsystem, etc., and the ground system could then command both the RHCP and the LHCP uplink phase shifters and VPDs to provide radiation patterns shaped in accordance with each of their needs, and independent of the needs of each other. Jammers may radiate randomly polarized signals and hence be detected by both JLE systems; still two JLEs are recommended to eliminate the possibility of a crosspolarized (to the JLE) jammer escaping detection.

2.2.5.2 Autonomous Operation

During the development of DSCS III the adaptive nulling algorithm was chosen for its straightforward, low-risk, low-power, low-weight, and ground-control characteristics. Since then the use of closed-loop nulling algorithms has matured. Virtually any one of these could replace the current system eliminating the JLE , ground-facility calculation, and uplink commanding. In short the current adaptive nulling algorithm could be replaced by an on-board closed-loop system that senses the presence of interfering signals and autonomously varies the antenna weights until the interference sensor indicates satisfactory suppression of the interfering signals. This process would have a relatively short-time constant (i.e., on the order of a few seconds) and it would automatically remove the null when the interfering signals vanish. Its adaption and

unadaption time constant differ by several orders of magnitude; consequently blinking jammers would not be effective.

Characteristic of all these nulling algorithms is the need to identify interfering signals. Assuming the interfering signal must occupy the entire operating bandwidth while the user's instantaneous signals, using an antijam waveform, occupy a fraction of the operating bandwidth (i.e., after despreading the received signals), interfering sources characteristically have greater EIRP than user terminals. Furthermore the user's signals have, or can have, some known characteristics such as their instantaneous frequency band. Using this information an algorithm can separate interference signals from user signals by setting the noise level, in a sensor circuit, higher than the user-signal level and lower than a troublesome interference signal level. This popular algorithm is called either a power inversion or Appelbaum-Howells algorithm. The former name implies high-power signals reduced below noise level in proportion to their strength. Operationally it is important they are reduced below the algorithm's noise level. With proper design this can result in up to 30-dB suppression of undesirable signals.

Two versions of the power inversion algorithm, applicable for both the RHCP and LHCP channels of the strawman dual-polarized system (Figure 1-2), are shown in Figure 2-5. Key to both these schemes is the need to make interference signals appear much larger than user signals. Thus a band reject filter and a dehopping mixer decrease the user's signal that is received by the correlator (Figure 2-5(a)) or the dither processor (Figure 2-5(b)). The correlator implementation (Figure 2-5(a)) collects a sample, through a directional coupler, of the operational bandwidth signal spectrum at the output port of each feed horn of the MBA. These ports could be either the RHCP or the LHCP ports. These signals are then

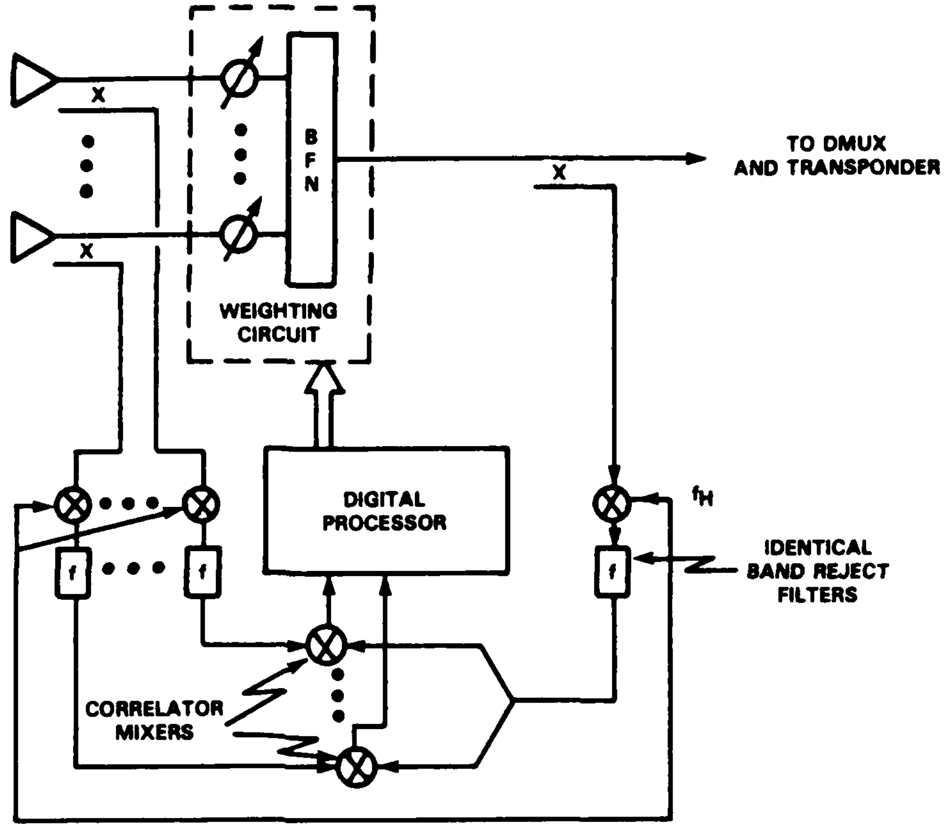



Figure 2-5(a). (U) Correlator Implementation

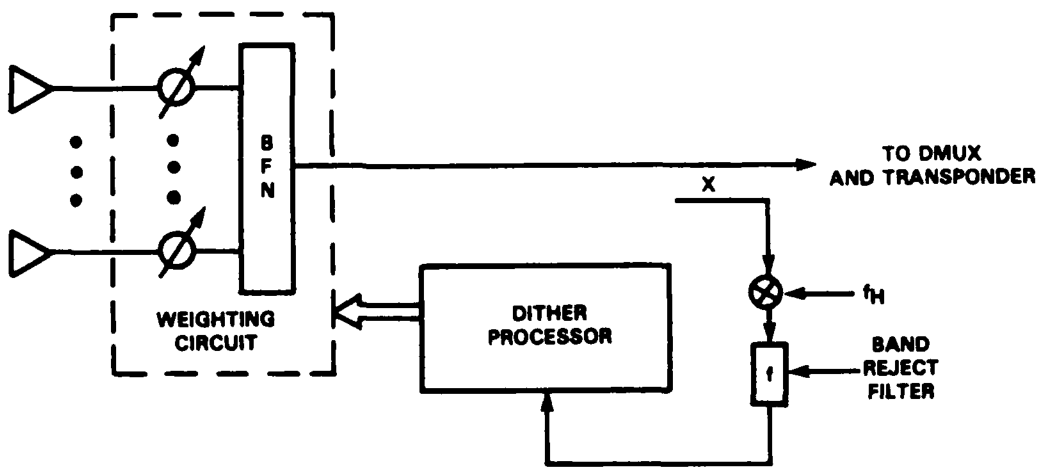



Figure 2-5(b). (U) Dither Implementation

**Figure 2-5. Comparison of Correlator and Dither Implementation**

mixed with a signal that is either a synchronized version of a pseudonoise (PN) or a frequency-hopping, antijam modulation on the received signals. Thus the antijam modulation is removed and the user signal occupies a known "nonvarying" bandwidth, whereas interfering signals are spread over the operational band. This mixer could also convert the incoming signals to an intermediate frequency (IF) band. The band reject filter should be capable of decreasing the level of user signals presented to the correlator more than 20 dB while attenuating the interference signals less than a few dB. An identical coupler-mixer-filter assembly provides the same modification of signals at the output of the BFN and present them to the correlator.

The correlator circuits are mixers that work in conjunction with a digital processor to derive control signals for adjusting the weight circuits. The weight circuit consists of the array of phase shifters and the BFN that can control the weight (vary phase and amplitude) of the signals received at each antenna port, sum them, and present this sum at the output of the BFN. Bandwidth of the correlator circuit would be equal to 360 MHz, the bandwidth of DSCS III channels 1 through 4. Correlation between any input signal and output signal results in changing the weight of the associated beam to decrease the correlation between input and output signals so that the jamming signal no longer appears in the output. Simultaneous adjustment of all weights, in the same manner, reduces the presence of interfering signals in the output. User signals are below the correlator noise level, consequently they are not sensed and hence are not removed from the output. This adaption process can occur in less than a second and will always provide radiation pattern "nulls" in the direction of interfering sources while maintaining the desired quiescent pattern in directions more than one-half power beamwidth (HPBW) from the direction of an interfering source. With some decrease in gain (approximately 10 dB) user terminals may be

within one-fourth of a HPBW of the interference source direction; this may permit reduced-rate jam-free communication that would be impossible with the current system.

The continuously adapting nature of this circuit permits satellite motion not possible with an open-loop algorithm. This feature will be discussed in more detail in Section 2.2.7.

It is important to note that Figure 2-5(a) shows an autonomous adaptive nulling system that contains all the features of a system designed for performance instead of for reduced cost, weight, power, and risk. For example, if adaption time can be increased to say 10 seconds, a single correlator could be shared among those beams in which a significant level of interfering signals is sensed. Several shared correlators could be used if the adaption time needed to be shorter. Adaption time should be limited so that natural, and/or operational, satellite motion does not significantly degrade the nulling performance of the system.

The dither implementation is inherently slower than the correlator implementation but it requires much less hardware and slightly more processing capability. The dither processor varies each weight, either sequentially, or in parallel, and derives a correlation between the variation of the weight and the output signal. This correlation is used to establish an optimum set of weights as developed in the circuit shown in Figure 2-5(a).

Note the directional couplers, mixer filters, and mixer correlators have been replaced by the dither processor; a significant reduction in costly, sensitive components. However, varying the weights to obtain adequate S/N to measure the correlation between input and output signal may result in varying the amplitude and/or phase of the communication signal and a concomitant degradation of the communication channel.

Little has been done to evaluate this effect and it is recommended that future study efforts examine this issue in greater detail.

Both nulling algorithms shown in Figure 2-5 use RF weights; that is the phase- and amplitude-modifying devices that operate in the frequency band 7900-8400 MHz. One could install a broadband low-noise amplifier (LNA) at the output port of the MBA. This would establish the uplink system S/N ratio and permit all beam-forming and nulling circuits to operate in "lossy" circuits and/or at a lower frequency. This implementation would require substantial development and study much of which could be derived from the Rome Air Development Center (RADC)-supervised EHF Satellite Adaptive Array Program (ESAAP) study conducted by General Electric at Valley Forge, Pennsylvania. The circuits discussed here were chosen primarily because they consist mainly of devices identical to or an extension of those used on DSCS III.

### 2.2.6 Incremental Reconfiguration

Circuits shown in Figure 2-5 would probably not use the set/reset sequence for changing the insertion phase and attenuation of the weight circuit. Rather the ferrite-phase shifters would be driven with pulses to incrementally increase or decrease the insertion phase. Periodically it may be necessary to set and reset all phase shifters (or perform a similar function) to ensure that incremental variations have not put their operating point near one edge of their operating range. Incremental variation of the weight circuit is desirable for adaptive nulling and reduction of stationkeeping. The latter will be discussed in Section 2.2.7.

2.2.6.1 Open Loop

Currently open-loop reconfiguration of the uplink MBA requires a large fraction of a minute. Most of this time is needed to pass commands through the decryption devices; the actual reconfiguration (i.e., driving the ferrites to obtain the desired insertion phase) requires about 100 microseconds. During the process of installing a new insertion phase it is necessary to drive the ferrite into saturation for the "set" stage and then drive it out of saturation a measured amount to complete the "reset" stage and install the desired insertion phase. During the set stage the insertion phase of a VPD or phase shifter can be in error by about 45 degrees. This error is compensated by the compensating phase shifter at the output port of the BFN. The amplitude error introduced by the Lincoln Laboratory switching algorithm described earlier can introduce up to a 3-dB decrease in the G/T or EIRP provided to a user by the satellite. Measured communication channel performance, obtained during development of DSCS III, proved this change in amplitude and phase did not increase the bit-error-rate (BER) or introduce errors in the transmitted bits.

During adaptive nulling, open-loop commands could be issued, with time tags, and stored on-board the satellite for appropriate reconfiguration of the uplink antenna to improve the null shape based on expected changes in the spacecraft attitude, etc. The time tags would enable a relatively fast succession of reconfiguration events as opposed to the current time to reconfigure.

By way of review, the Lincoln Laboratory switching algorithm switches all phase shifters in the VPDs, at a given level in the BFN, at the same time. First those in level one (i.e., closest to the antenna feed horns) are switched, then the second level, and so on to the sixth level or output VPD. Each VPD is switched to its new state if the power to a user in

both the old and the new states is not decreased by more than 3 dB; otherwise it is set to provide equal-power division. This is true for all levels on the first pass except the sixth level or output VPD; this VPD is switched to its final power division state during the first pass. After setting the sixth-level VPD, the fifth-level VPDs are set to their final power division state. Next, the fourth level and so on until all VPDs are in their final state. Then all 61 phase shifters are set to their final state. While switching each level, the VPDs are first driven to an equal power division state. This introduces an approximate 45-degree insertion phase error with respect to the insertion phase of these devices when they are set to give the desired power division. A one-bit phase shifter located at the output of the sixth-level VPD is switched during this "set" switching stage of each VPD. The combined insertion phase, of the one-bit phase shifter and a level of VPDs in the "set" stage, is less than a few, possibly one, degrees.

With incremental switching, gradual reconfiguration of the antenna radiation pattern is possible; it may not be necessary to have a one-bit compensating phase shifter and the pattern might be made to change shape gradually in accordance with a prescribed manner. However, it will be necessary to periodically reconfigure, using the current algorithm because incremental reconfiguration does not permit accurate knowledge of the final weights installed. This can be done only through the "set/reset" method.

2.2.6.2 Autonomous

Incremental reconfiguration can be useful if:

1. There is a feedback sensor measuring the performance of the reconfigured system.

2. There exists a proven succession of incremental changes that produce the desired result.

The first condition can be satisfied by an on-board autonomous nulling algorithm. In fact, it is entirely likely autonomous adaptive nulling will use incremental reconfiguration techniques. The second condition can most likely be satisfied but study is required.

### 2.2.7 North-South Stationkeeping

Satellites in an equatorial 24-hour orbit appear stationary to an earth terminal. However, various environmental, launch, and weight constraints often dictate orbits with a slight inclination for the synchronous satellite as in the case of DSCS III. Consequently the satellite's nadir describes a "figure-of-eight pattern" on the earth's surface. This causes the elevation and azimuth angles measured by a user terminal to vary slightly at a diurnal rate. In addition the satellite's attitude, with respect to the earth, changes giving rise to errors in pointing its reference toward the center of the earth. Solar forces tend to force the orbit of the DSCS III satellite to precess. The greater the precession, the greater attitude correction is required to maintain the desired attitude of the satellite. This attitude can be maintained by appropriate loading and unloading of the satellite's inertia wheel. It is also possible to periodically correct the satellite's orbit, maintaining it at a low-inclination angle by expending fuel. The DSCS III satellite carries approximately 400 lbs of fuel for this purpose. This was chosen because unloading and loading the inertia wheel may upset the earth sensor and/or sun sensor introducing unpredictably large pointing errors; whereas, accurate pointing is required to maintain a pattern null in the desired direction. Clearly this is a complicated problem which can be solved by either attitude control or antenna pattern reconfiguration and thus eliminate

the need for NSSK propellant and permit much valuable communication capability to be added.

In other words, eliminating NSSK propellant will allow the satellite's orbit to precess giving rise to a diurnal variation in the satellite's nadir trace on the earth. If this variation in pointing is corrected by reconfiguring the antenna, there is the risk each reconfiguration will introduce errors in the bit stream during the reconfiguration transients. However, by incrementally reconfiguring the antenna pattern it may be possible to change the antenna pattern without introducing any bit-error producing transients. Major orbit precession could be corrected by NSSK-propellant control but with much less propellant required for the satellite's mission. Hence with frequent incremental radiation-pattern reconfiguration and much less frequent NSSK correction, it may be possible to use the concomitant weight saving to add communication capability, perhaps EHF communication capability. A detailed study of this is beyond the scope of this study but the subject is worthy of review and further study.

#### 2.2.7.1 Open Loop

Incremental reconfiguration could be carried out by periodic control from the earth using the algorithm currently used to calculate the reconfiguration weights. Some changes in the ferrite-phase shifter driver would be required; specifically, the "set" pulse would have to be eliminated and the "reset" pulse modified substantially so only a relatively small incremental change in the insertion phase would be made during a typical incremental reconfiguration. It would be necessary perhaps through experience to occasionally reconfigure the antenna pattern using the set/reset pulses.

During the development of DSCS III communication, signals were passed through a switching VPD (i.e., with set/reset pulses) with no noticeable increase in BER. These tests were performed at Fort Monmouth; the test data and results should be reviewed. Present and future plans can be formulated with much less risk, if this or similar data is made available. It may even be possible to ascertain the sensitivity of DSCS III to reconfiguration transients.

2.2.7.2 Autonomous

With autonomous control of the antenna's radiation pattern it may be possible to eliminate NSSK completely. A beacon signal transmitted toward the satellite could provide a "pointing reference" for incremental reconfiguration in the absence of an interfering source; in the presence of an interfering signal, the adaptive nulling algorithm would continuously point a null in the desired direction while maintaining a best fit to the desired quiescent pattern. Indeed the saving in NSSK-propellant weight would more than permit the addition of an autonomous nulling algorithm.

2.2.8 Acquisition and Synchronization Considerations

Current DSCS III communication channels must operate in the presence of noise and interfering signals peculiar to a single-polarized system. In particular, signals radiated by friendly users are at a different frequency and must be filtered by a frequency demultiplexer and other filtering devices. A dual-polarized system such as the strawman system described in Figure 1-2 and the foregoing permits user terminals to radiate signals at the same frequency: only polarization provides appropriate isolation. However, polarization impurity as described in the foregoing sections permits friendly user signals to interfere with a user's signals at the same frequency but operating on the crosspolarized channel.

This self-jamming effect degrades communications and makes acquisition and synchronization with the transmitted bit stream more difficult than with a system such as DSCS III. This section discusses acquisition and synchronization function when a dual-polarized system is operated with open-loop and autonomous-antenna reconfiguration.

#### 2.2.8.1 Open Loop

During acquisition a system's S/N is degraded and usually the instantaneous bandwidth is narrower than after acquisition has been completed. The reduced S/N renders the system more vulnerable to disruption by self-jamming crosspolarized signals and transients introduced by reconfiguring the antenna. These same noise-like sources can affect the maintenance of bit-stream synchronization.

Since open-loop reconfiguration can be scheduled to occur under favorable conditions and/or disciplined operation of the user terminals can be enforced by the ground controller; it is unlikely either the adaptive-nulling algorithm or the incremental reconfiguration will seriously degrade normal acquisition and synchronization processes.

#### 2.2.8.2 Autonomous

Autonomous adaptive nulling or reconfiguration (e.g., to reduce the need for NSSK) unless properly constrained might attempt a reconfiguration when a terminal is in the acquisition phase. That is, one channel could suddenly be subjected to a jamming attack forcing the adaptive-nulling algorithm to begin reconfiguring the antenna. A user on another channel supported by the uplink MBA might be in the acquisition process. The reconfiguration process might introduce unexpected noise bursts making acquisition unusually difficult. If the adaption process is delayed long enough for the terminal to acquire the

uplink, reconfiguration to suppress an interfering source could be carried out without disrupting communication on other channels. Alternately when this rare event occurs the user could attempt to acquire again.

A detailed study of the acquisition process for autonomous nulling is beyond the scope of this report. In addition, use of dual-polarization techniques would not significantly affect the acquisition process although some additional functions may be required.

## 2.3 GENERAL PERFORMANCE REQUIREMENTS

Successful operation of DSCS III with a dual-polarized system depends largely on the degree of channel isolation that can be received. The effects of polarization purity and match on channel isolation were covered in detail in Section 2.2.4. These are summarized briefly here to emphasize their importance.

### 2.3.1 Polarization Purity

In Section 2.2.4 it was pointed out that circularly polarized antennas usually have an axial ratio greater than 1 dB, but perhaps less than 2 dB. Perfect CP is virtually impossible to produce. Furthermore the polarization match between the terminal antenna and the spacecraft antenna is almost never perfect. Even if they were matched, rain and/or ice along the propagation path changes the polarization of a wave propagating through it (this is discussed in Section 2.5.1). These three factors combine to produce crosscoupling between crosspolarized channels. In order to minimize this degree of coupling, concerted effort should be made to achieve nearly CP of the antennas involved. If the terminal employs a polarization tracker (see Section 2.4.2.3) it would be possible to reduce the depolarization effects of the propagation path. Polarization of the uplink and downlink antennas should have an

axial ratio less than 0.5 db if possible. Then the terminal could infer the depolarization on the uplink and by measuring depolarization of the downlink signals. The polarization tracker can be designed to receive and transmit orthogonally polarized signals and thereby, in principle, transmit a wave polarized to compensate for depolarization effects due to the propagation path.

One method of assuring adequate polarization match is to specify antenna-polarization axial ratio less than 0.5 dB. It will be undoubtedly difficult to achieve this, however, meeting this performance specification will guarantee less than -30-dB crosspolarization coupling between channels operating at the same frequency and in clear weather.

### 2.3.2 Polarization Match

It is important that the polarization of the spacecraft uplink (downlink) antenna match the polarization of the terminal's uplink (downlink) antenna when the propagation path does not alter the polarization of the wave propagating along it. Since the latter can occur it may be necessary to adjust the terminal antenna's polarization to compensate for the effect of the propagation path. This polarization-matching function can be readily implemented on the downlink (see Section 2.4.2.3) because the terminal can vary its polarization to maximize the received signal. It is not possible for the satellite to carry out the same function for at least two reasons:

1. The satellite antenna would have to match its polarization to all uplink terminals accessing the satellite at a given time. This is impossible.

2. The circuit required to adjust the polarization of the satellite MBAs would be complex, heavy and costly, and possibly impractical.

The system should plan to use terminal antennas capable of adapting their polarization to match that of the incoming wave. This information should be used to alter the polarization of the terminal's uplink antenna to provide the best polarization match possible to the satellite's uplink antenna. It may be possible to sense the degree of mismatch on board the spacecraft and alter the terminal's uplink antenna accordingly. It may also prove unnecessary to do this except in severe rain storms and then attenuation may cause a more serious disruption of the link. These issues will be discussed in Sections 2.5.1 and 2.5.2.

### 2.3.3 Channel Isolation

Channel isolation effects were addressed in Sections 2.2.3 and 2.2.4 and the results are indicated in Figure 2-4. In order to limit channel degradation to less than 1 dB, channel isolation must be equal to or less than the channel $S/N_oB$ (i.e., the S/N ratio in the absence of crosscoupled signals). If the terminals interfering with one another illuminate the satellite with unequal power flux, the channel coupling due to crosspolarization may have to be even less--this may be necessary in the case of terminals with unequal EIRP and/or unequal path attenuation.

As in the case of DSCS III channel demultiplexing and multiplexing, filters must provide adequate frequency filtering. It may be necessary to increase channel isolation by increasing the frequency filtering between channels. On the other hand, no degree of frequency filtering will prevent coupling between channels operating over the same frequency band.

## 2.4 TERMINAL CONSIDERATIONS

DSCS III terminals are currently designed to transmit RHCP signals and receive LHCP signals. If the payload is upgraded to dual-polarization operation, the terminals will have to be upgraded to use the additional "dual-polarized" channels. Define the current polarization of both terminal and payload antennas as copolarized channels and the upgrade dual-polarized channels as crosspolarized channels. Some of the payload antennas must be upgraded to handle both copolarized and crosspolarized signals; terminals can remain as they are (i.e., with copolarized capability), can switch to crosspolarized, or add crosspolarized. This section addresses the options, their expected performance, and associated special characteristics.

### 2.4.1 <u>Current Configuration (General)</u>

Although there are many varied DSCS III terminals, the RF system of at least a typical terminal, and perhaps of all terminals, can be represented by the block diagram shown in Figure 1-3(a). The terminal's single antenna is a paraboloid reflector, from four to sixty feet in diameter, illuminated by a dual-polarized feed. The latter is represented as a "dual-polarized antenna port" in Figure 1-3. The current design uses the natural isolation (approximately 30 dB) between orthogonal-polarization ports and bandpass filters in the transmit and receive bands, represented by FT and FR, respectively, in Figures 1-3 and 2-6. LHCP-received signals are usually amplified by a LNA, demodulated, and delivered to the user. The user can initiate signals that are modulated, by the modem, and amplified and transmitted as RHCP signals.

#### 2.4.1.2 Diplexer Configuration

The diplexing filters are characteristically bandpass devices centered on the transmit (7250-7750 MHz) and receive

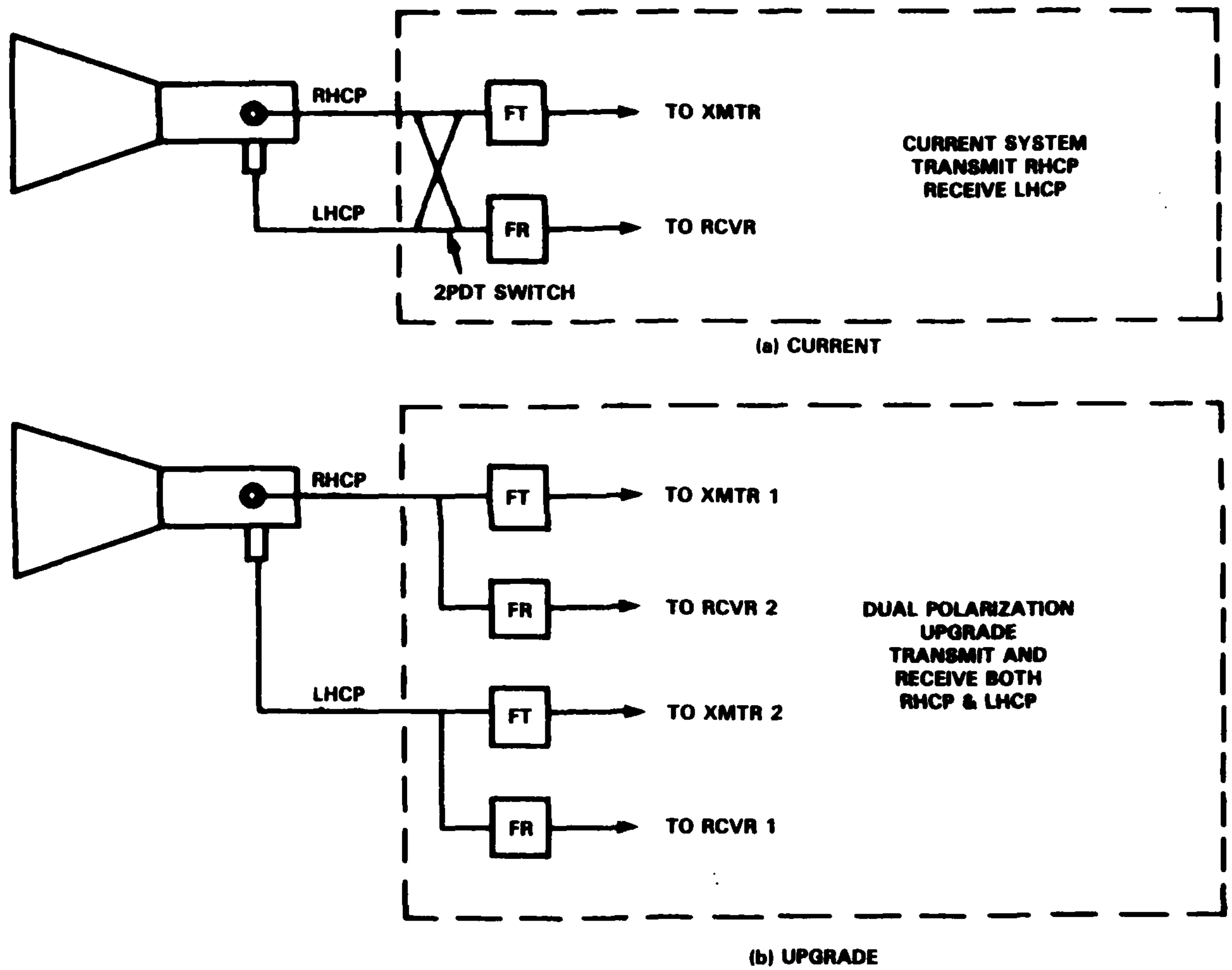



**Figure 2-6. Typical Dual-Polarized Terminal RF Without Adaptive Polarizer**

511012.0

(7900-8400 MHz) bands. An additional band-reject filter (i.e., included in FT) is sometimes included in the transmit filter to provide additional rejection of transmitter signals in the receive-frequency band. The inherent isolation between LHCP and RHCP antenna ports reduces the required power handling capability of receiving diplexer filters substantially. This inherent isolation exists regardless of the purity of polarization. Since only copolarized channels are used, polarization purity is not necessary for isolation between channels; it is required only to reduce loss of copolarized signals. Specifically the signal power $P_s$ received by an antenna with polarization axial ratio $AR_a$, when it is illuminated by a wave with polarization axial ratio $AR_w$, is given by (References 2 and 3):

$$P_s/P_o = 1/2(1+((AR_w^2-1)(AR_a^2-1)\cos 2A \pm 4AR_wAR_a)/D) \tag{9}$$

where

$$D = (AR_a^2+1)(AR_w^2+1),$$

and A is the angle between the major axes of the polarization ellipses; the minus sign is used if the waves are of opposite sense rotation, and $P_o$ is the power at the antenna's output port when the polarization of the wave is the same as the polarization of the antenna.

If either $AR_a$, or $AR_w = 1$,

$$P_s/P_o = 1/2[(1+2AR/(AR^2+1)]. \tag{10}$$

From (9) the maximum reduction in $P_s/P_o$ occurs when A = 90 degrees (i.e., the major axes of the polarization ellipses are perpendicular to each other) and

$$P_s/P_o = 1/2(1-((AR_w^2-1)(AR_a^2-1)-4AR_wAR_a)/D). \quad (11)$$

Using (11), power loss ($10\log P_s/P_o$) was calculated as a function of $AR_a$ for various values of $AR_w$; the results are plotted in Figure 2-7. Note if both $AR_a$ and $AR_w$ are less than 1 dB, the worst-case mismatch introduces less than 0.25 dB.

### 2.4.2 Candidate Dual-Polarized Systems

Modification of a terminal, or design of a new terminal to operate with both the crosspolarized and copolarized channels could provide backward compatibility for improved flexibility. If economy rules, it is possible to modify existing terminals permitting them to operate with either crosspolarized or copolarized channels but not both. The following section presents the salient details that should be considered.

#### 2.4.2.1 Transmit and Receive Opposite Sense CP Signals

The present DSCS III terminals transmit RHCP and receive LHCP signals. Since they use a single antenna for both functions, diplexing filters must be used. As indicated in Figure 2-6(a), the inherent isolation between RHCP and LHCP antenna ports is included in the diplexer design; that is, the total isolation between transmitter and receiver is the sum of the attenuation of either FT, or FR, and the isolation between the RHCP and LHCP antenna ports. The latter is usually more than 30 dB. Including a double-pole, double-throw switch permits transmission of either RHCP or LHCP signals, while receiving the opposite sense polarization.

#### 2.4.2.2 Transmit and Receive Same Sense CP Signals

Some terminals may be required to access both copolarized and crosspolarized channels simultaneously. This requires the transmission and reception of signals with the same polarization.

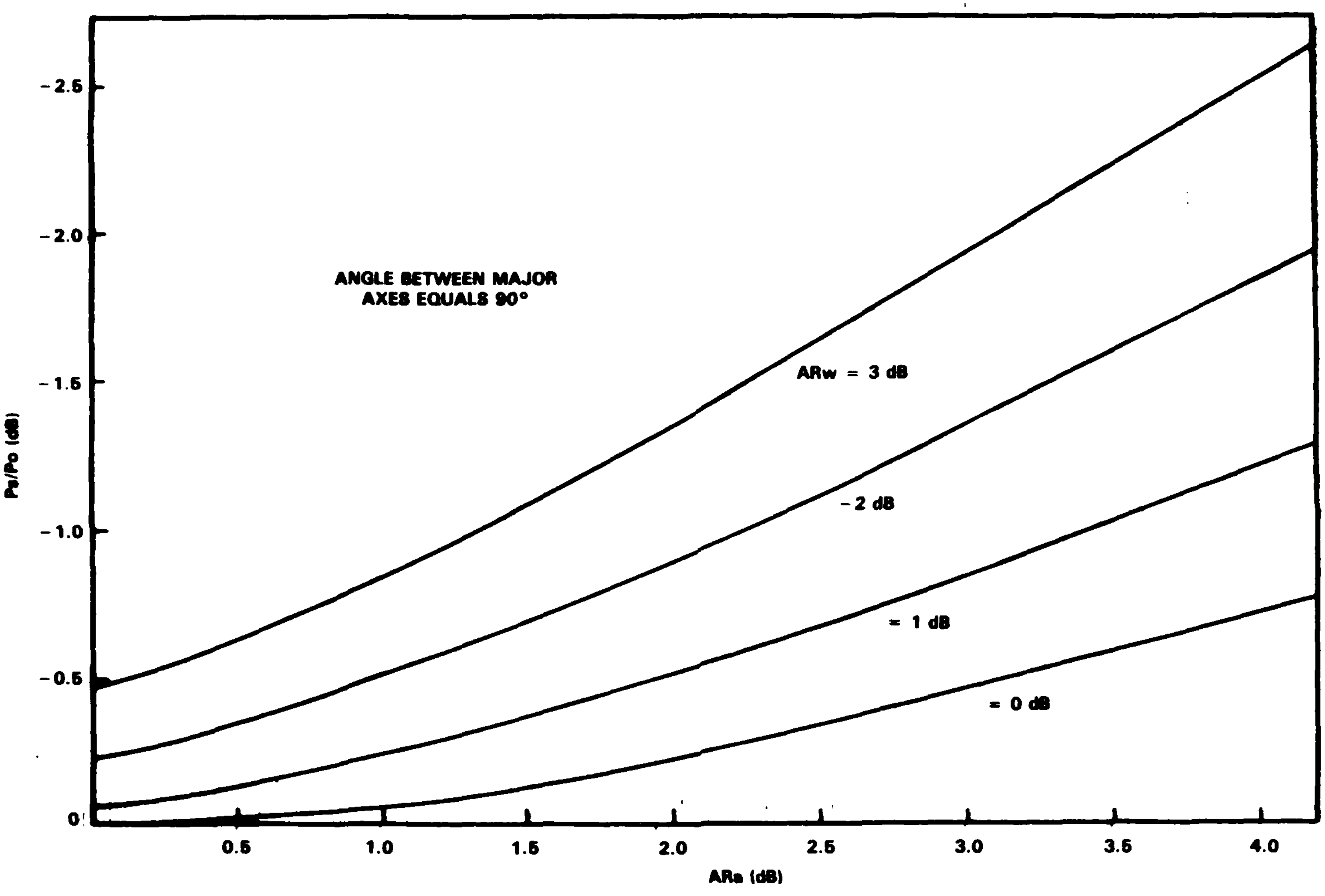



**Figure 2-7. Loss Due to Polarization Mismatch**

511013.0

Consequently, the diplexing filters must be designed to provide the total isolation required between the transmitter and the receiver; that is, at least 30 dB more than in the present system. This arrangement is shown in Figure 2-6(b) where FT and FR are both connected to the copolarized "antenna" port.

2.4.2.3 Track and Adapt to Polarization of Received Signals

It was pointed out in previous Sections (2.2.3 and 2.2.4) that polarization purity is essential to reduced self-jamming due to cross-channel coupling. It is possible to design the terminal and payload antennas such that axial ratio is less than 1 dB and guarantee channel isolation greater than 30 dB. However, depolarization due to rain and/or ice along the propagation path can reduce copolarized-to-crosspolarized channel isolation to 15 dB. This effect can be compensated for on the downlink by including a polarization-adaption circuit in the terminal's antenna system. This circuit (see Figure 1-3(b)) adjusts the relative phase and amplitude of signals received at the RHCP and LHCP antenna ports so that the signal at the crosspolarized output is minimized. This automatically maximizes the copolarized signal received at the copolarized antenna port and reduces cross channel coupling.

The adaptive polarizer uses a ferrite-variable phase shifter for low loss and a variable-power divider similar to that used in the payload. The RHCP and LHCP signals are correlated with the crosspolarized signals. The result is used to drive phase shifters P1, P2, and P3 so that the signals become uncorrelated; i.e., signals sensed by the correlators vanish from the crosspolarized port. It is assumed the desired incident signals will be much stronger (greater than 10 dB) than the undesired signals, and the correlator noise level is intentionally greater than the undesired signal level. Once adaption is achieved, the copolarized and crosspolarized signals will be approximately the same as if there were no

polarization changes due to the propagation path. Once the polarization adaption is achieved, the adaptive polarizer will track only changes in the polarization of the incident wave.

Unfortunately the satellite antennas cannot use an adaptive polarizer because they must serve several terminals at the same time and each terminal will illuminate the satellite with a slightly different polarization. The difference is due to their individual construction, the propagation path, and their location on the earth.

2.4.2.4 Polarization Purity

As pointed out previously, mismatch between the polarization of the wave incident on the payload antenna and the antenna's polarization causes undesirable coupling between the copolarized and crosspolarized channels. This mismatch can be minimized if the polarization of the terminal's antenna is as circular as possible. It will be difficult to improve the axial ratio of terminals already operational and not designated for an upgrade. However, all terminals designated for operation in the crosspolarized channels must have a polarization AR less than 0.5 dB to prevent them from jamming the existing DSCS III terminals. It would prove unfortunate if new crosspolarized terminals literally jammed existing copolarized terminals because of intolerable cross-channel coupling due to a high axial ratio.

2.4.2.5 Polarization Match

Satellite uplink and downlink antennas can achieve polarization "match" only through careful design. That is, uplink antennas can be orthogonally polarized to the downlink antennas. This is difficult to achieve in two different antennas operating over a different frequency band. Nevertheless, purity of polarization implies polarization orthogonality between uplink and downlink antennas since

terminal antennas can readily achieve a high degree of polarization orthogonality because they transmit and receive with the same antenna.

#### 2.4.2.6 Channel Isolation

It will probably be rather simple for the terminal to achieve good isolation between copolarized and crosspolarized signals on the downlink; particularly if an adaptive polarizer is used. It probably will be much more difficult for the terminal to choose the correct polarization of its radiated signals unless the downlink and uplink propagation effects are nearly the same. This is discussed further in Section 2.5.2.

## 2.5 SYSTEM CONSIDERATIONS

Many elements of systemic, as opposed to component, influence were discussed in the previous sections. In this section important systemic effects not mentioned, or fully discussed, previously are addressed.

### 2.5.1 Propagation Effects

#### 2.5.1.1 Rain

Electromagnetic waves propagating through earth's atmosphere encounter attenuation due to particulate scattering and molecular absorption. These waves also undergo a delay that can be accounted for by an insertion-phase variation. These effects are usually small (less than 1 dB and 1 degree) at these frequencies, when the path is through "clear air"; that is, no clouds, rain, dust, etc., are along the terminal-satellite path. Rain is the principal cause of attenuation and it is generally polarization independent and can introduce up to about 5-dB attenuation 0.1 percent of the time [Reference 5] when the terminal is located in a heavy rainfall area such as

Washington, D.C., or Florida. Rain attenuation-induced fades are, in general, within the design margin of most DSCS III terminal-satellite links, have the same effect on both copolarized and crosspolarized channels, and therefore, require no systemic consideration than for the current DSCS III.

Vertically polarized signals experience a delay when propagating through rain that is, in general, different than for horizontally polarized signals. This differential delay changes the polarization of elliptically polarized waves, or linearly polarized waves, that are neither vertically or horizontally polarized. It is nonspherical rain drops that cause this differential delay, and the difference in their canting angle (i.e., the orientation of their major axis) tends to average this delay effort over the volume of the earth-satellite path. It follows that vertically or horizontally polarized signals are not depolarized when propagating through rain (see Reference 5). Measured data indicates that heavy rain can introduce up to -15-dB crosscoupling between orthogonally polarized signals. Crosscoupling is defined as the ratio of crosspolarized to copolarized signals as the wave exits the rain-filled medium. This -15-dB crosscoupling occurs simultaneously with about 10-dB attenuation neither of which occurs more than 0.17 percent of the time. It is also likely that 10-dB attenuation would cause an unacceptable increase in the BER regardless of the added noise due to crosscoupling between dual-polarized channels.

#### 2.5.1.2 Ice

Ice in the form of frozen droplets does not introduce significant attenuation of an electromagnetic wave propagating through it, but it does introduce insertion-phase shift that is polarization dependent. Ice in this form occurs most frequently at altitudes above the melting layer in the earth's atmosphere and usually in the tropic climates. It is believed

to be the dominant cause of "rain-induced" depolarization. This latter statement is based on measured depolarization in light rain, when the earth-satellite path intersected the melting layer of the local rainstorm. Depolarization or attenuation, caused by either rain or ice, is of little consolation if they cause an intolerable increase in the link BER. The specific cause and effect are pointed out primarily to indicate the possible occurrence of significant depolarization with moderate attenuation. An indepth analysis of this effect is beyond the scope of this report.

### 2.5.2 Uplink/Downlink Polarization Correlation

These ice and rain depolarization effects are only slightly dependent on frequency. Hence, measurement of the polarization of a wave received on the downlink can be used to determine the polarization of the wave transmitted on the uplink that will arrive at the satellite appropriately polarized to reduce coupling between crosspolarized channels. The polarization tracker described in Section 2.4.2.3 usually has an "orthogonal" port which converts input (transmit) signals to waves orthogonally polarized to the waves received from the satellite; i.e., a polarization that may be different from that of the satellite's uplink antennas. Propagation through the same rain (i.e., as experienced by the waves on the downlink) tends to remove the depolarizing effect introduced on the downlink. This results in a wave, incident on the satellite antennas, that is polarized almost the same as with no rain along the path.

In summary, DSCS III CP waves are depolarized on the downlink if they pass through rain. A polarization tracking receiver adjusts the terminal's antenna polarization to match the polarization of the received signal. At the same time it adjusts the transmitted signal's polarization so that the wave arriving at the satellite is nearly polarization matched to

that of the satellite's uplink antenna. Perfect match is not possible because the up and downlink waves have a different wavelength (i.e., their interaction with the raindrops is slightly different) and the satellite's up and downlink antennas are not "orthogonally" polarized to the same accuracy as the terminal's antenna. This is because the terminal uses the same antenna and feed for both transmitting and receiving signals, whereas, the satellite uses different antennas for this function.

### 2.5.3 Uplink Power Control Considerations

Transponder satellites, such as DSCS III, can be modeled as a linear amplifier that receives signals from the earth, amplifies them with very slight distortion, adds some noise, and transmits them back to earth. This model, and the satellite, allocates the downlink EIRP in accordance with the amplitude of the received (i.e., from the earth) signals. As long as the input signals do not saturate the satellite transmitter, the larger amplitude signals are allocated a larger portion of the downlink EIRP than is allocated for the smaller amplitude signals. This is in accordance with system design as long as the terminals are appropriately disciplined and transmit no more than the EIRP assigned to them. When this discipline fails and a terminal increases its EIRP above the assigned level, terminals with lower EIRP will lose some of their allocated downlink EIRP. In effect, the lower EIRP terminals experience a decrease in the "gain" of the satellite transponder. This effect is called "power robbing" and is eliminated only through discipline among the terminals.

In a dual-polarized configuration, of the type shown in Figure 1-2, coupled power between crosspolarized channels appears as noise in the crosspolarized channel and also, in effect, robs some of the downlink power. Appropriate net discipline might alleviate this effect or in some cases

eliminate it. A suitable algorithm and enforcement activity are beyond the scope of this report but perhaps useful to the architectural design of a dual-polarized system.

## 2.6 ESTIMATED SATELLITE WEIGHT AND POWER IMPACT

A detailed weight and power estimate requires a reasonably firm description of the proposed dual-polarization configuration. A rough estimate can be derived from proposed configuration (Figure 1-2) and a general algorithm that estimates the spacecraft weight and power required to support the estimated payload weight and power.

An estimated weight and power budget for the additional crosspolarized channels, based on general knowledge, is given in Table 1-2. This weight and power must be added to the current payload weight and power to obtain an estimate of the total weight and power of a dual-polarized payload. A more accurate estimate could be derived using known weight and required power of DSCS III components. This increase in accuracy would be misleading until a more detailed design of the dual-polarized payload is available. The values given previously in Table 1-2 are representative (probably within 25 percent) of a specific configuration designed to provide essentially the same communication capacity.

The spacecraft launch weight and power required to support the additional crosspolarized channels can be estimated using

$$W_s = 1.7(W_p + .6P_p) \tag{12}$$

where $W_s$ = satellite weight,
$W_p$ = payload weight,
$P_p$ = payload power.

Using (12) and Table 1-2 the estimated launch weight of the crosspolarized channels is 630 lbs. Assuming power regulation, etc., will be required, the spacecraft would have to supply 300/.85 = 350 watts to the "crosspolarized" payload; i.e., this power is in addition to that required by the current DSCS III payload.

The reader is reminded the launch weight estimate is based on (12) which was not derived from modification of spacecraft data. Rather it was derived from a large number of operating spacecraft.

# CHAPTER 3

# ANTENNA DESIGN OF A WIDEBAND EHF PACKAGE (PAYLOAD)

Chapter 2 addressed the addition of crosspolarized channels to increase the communication capacity of the DSCS III payload. Increasing antijam protection through spatial discrimination was considered as an extension of the DSCS III payload capability to the crosspolarized channels and in the form of an on-board adaptive-nulling algorithm for the uplink MBA. Antijam protection could be improved by increasing the size of the MBA and/or using an on-board adaptive-nulling algorithm. However, an additional EHF package could provide a somewhat larger increase in communication capacity and a substantial increase in antijam protection through increased spatial discrimination and spread-spectrum modulation. This section addresses communication subsystem and signal-processor characteristics of a candidate EHF payload designed to serve users requiring wideband communication capacity. The signal processor and spread-spectrum waveform interact significantly with the antenna design; consequently both must be considered in order to place the appropriate tradeoff or compromises in evidence.

User coverage requirements strongly influence the antenna design making it important to define at least two general types of coverage, namely, earth FOV and area FOV. This leads to a discussion of the interdependence of antenna size, jammer-user separation, and jam-free, or quiescent, antenna radiation pattern. This relationship is presented first followed by a description of a candidate antenna, an estimate of its performance characteristics, and impact on the spacecraft weight and power requirements.

## 3.1 COVERAGE DEFINITION

Expected location of user terminals determines the radiation pattern required to service them. Uncertain location requires an earth-coverage antenna pattern. Small communities such as those that might be called a battle group (Naval armada), battle theater (Army engagement), etc., require coverage over an area ranging from a few tens to a few hundreds of miles in diameter. Point-to-point communication links generally desire one or more narrow antenna beams pointing in the direction of each user terminal when it is transmitting signals to, or receiving signals from, the spacecraft. In all cases the terminals must point their antenna beams toward the spacecraft. In the interest of completeness, and to establish some fundamental "coverage" rules, the following section discusses these radiation coverage patterns in some detail.

### 3.1.1 Earth FOV

Satellites in a synchronous orbit are approximately 22,300 statute miles above the surface of the earth. At that altitude the earth subtends an angle of 17.45 degrees. At lower altitudes this subtended angle is larger as indicated in Figure 3-1 where the subtended angle is shown for altitudes between 200 miles and 22,300 miles.

It is important to note that defining the coverage pattern specifies the maximum gain of an antenna designed to provide that coverage. Assuming a circular coverage area with diameter $\theta_o$, $G_{max}$ the maximum gain is given by:

$$G_{max} = 4\pi(57.2)^2 \approx 41,000/\theta_o^2. \qquad (13)$$

For an earth-coverage antenna pattern $\theta_o = 17.45$ and $G_{max} = 21.32$ dB; (13) also indicates the gain of an antenna varies inversely as the square of the coverage area's angular

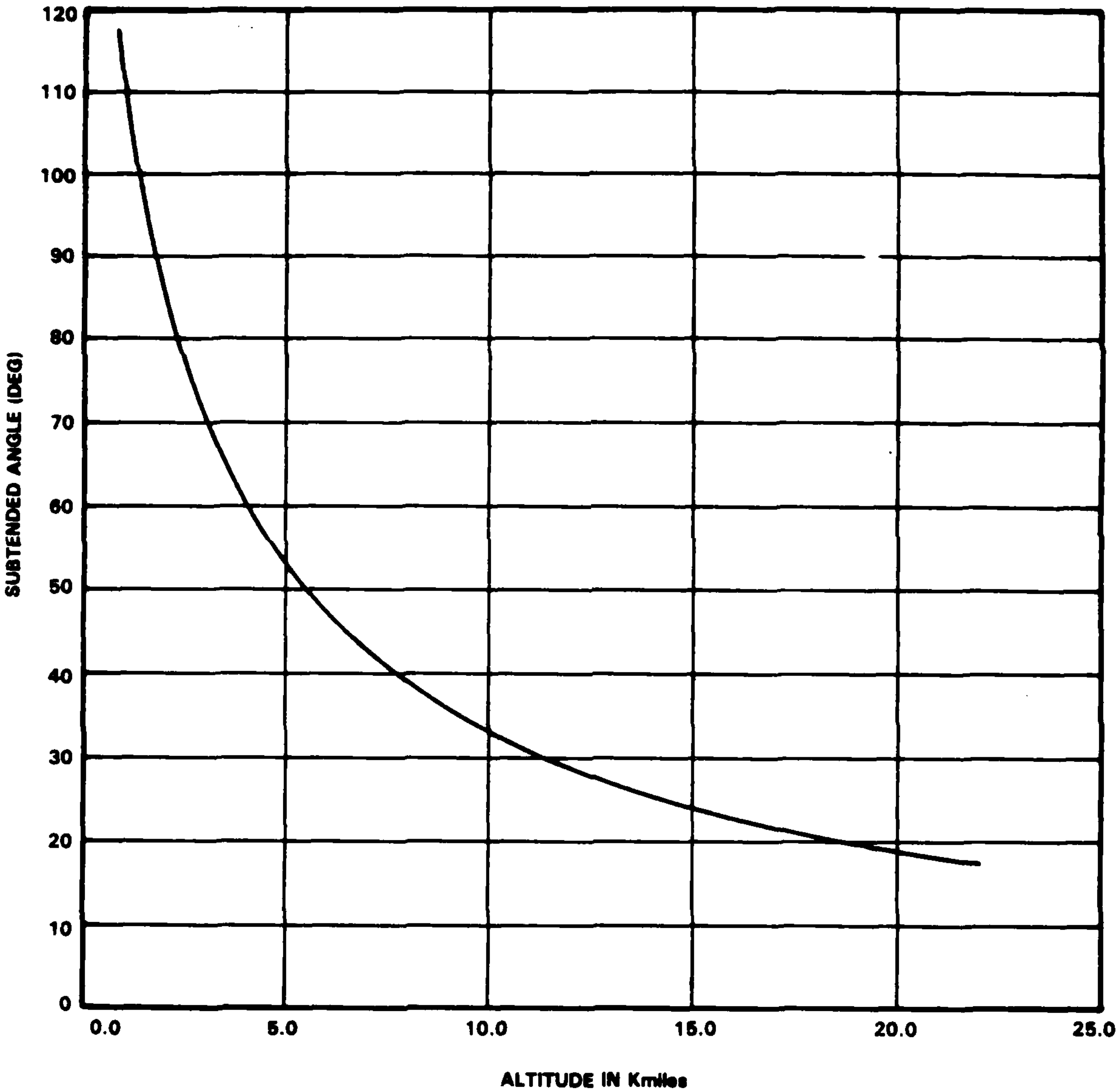



**Figure 3-1. Earth Subtended Angle**

511014.0

diameter. Although most extant earth-coverage antennas on synchronous satellites have less than 17.5-dB gain over the earth's disk, it is possible to design antennas that provide greater than 17.5-dB gain over the earth's disk. The DSCS III MBAs provide greater than 18-dB gain over earth's disk when the weights in the beam-forming network are adjusted appropriately. However, the aperture of these MBAs is much larger than that of the earth-coverage horns, operating over the same frequency band, on the same spacecraft; thus the price for increased gain is complexity and antenna aperture size. Operation at EHF introduces the possibility of using electrically large aperture antennas whose radiation pattern can be shaped to realize greater than 20-dB gain over the earth's disk. Detail design and or analysis of this type antenna is outside the scope of this report; nevertheless, antennas of this type can be important to the development of an EHF MILSATCOM system.

### 3.1.2 Area Coverage

Satellite antennas designed to cover less than the earth's disk are often referred to as "area-coverage antennas." The angular diameter of an area served by an area-coverage antenna usually varies from about 1 to 15 degrees. In this report, area coverage implies coverage of a circular area about 1 to 2 degrees in diameter. The term "spot-beam" coverage is used to indicate a conventional paraboloid antenna with a single beam or excitation of a single beam of an MBA.

It is important to point out that all antenna radiation patterns do not provide maximum gain over the coverage area when their aperture size is chosen to make the half-power beamwidth equal to the angular diameter of the coverage area. The conventional paraboloid provides maximum gain over a coverage area of angular diameter $\theta_C$ when the 4.3-dB beamwidth of its radiation pattern equals $\theta_C$. Oversize aperture antennas (such as an MBA) maximize gain over their

coverage area when $\theta_c$ is less than the half-power beamwidth. In other words, flat-nosed beams have steeper sides and therefore provide more nearly uniform gain over the coverage area; somewhat like that provided by the beam assumed to derive (13).

## 3.2 NULLING RESOLUTION

This subject was addressed in a recent report [Reference 9]. A brief review and further refinement of the analysis are presented here resulting in slightly more accurate results that indicate the dependence of minimum jammer-user separation as a function of antenna-aperture and quiescent-radiation pattern.

Many previous studies indicate the minimum user-jammer separation that can be tolerated is determined by the antenna aperture's largest dimension. Characteristically a TPA is more economical than an MBA when both are designed to achieve the same resolution. On the other hand, the TPA is characteristically more vulnerable to jammers located a long distance from the user. Consequently a TPA of N elements can be disrupted, perhaps completely disabled, by a few more than N jammers located anywhere in the FOV of an element of the array. In contrast, the MBA cannot be effectively jammed by any number (i.e., less than say 100) of jammers located outside the instantaneous FOV of the MBA. In this section the potential of combining the better features of these fundamental antennas is addressed.

### 3.2.1 Aperture Dependence

It is universally agreed that adaptive-nulling processors, in concept, produce a "null" in the direction of a jammer by forming a maximum directivity beam in the direction of the jammer and subtracting it from the quiescent or jam-free antenna pattern. A maximum directivity beam is produced by a

uniform aperture distribution; the subtraction process assumes equal gain of both the quiescent and maximum directivity beams in the direction of the jammer. Increasing the antenna aperture decreases the maximum directivity beam's half-power width proportionately. This in turn results in a proportional decrease in the angular width of the null and the tolerable jammer-user separation. Decreasing the antenna size increases the tolerable jammer-user separation. Nulling resolution or minimum jammer-user separation are dependent on the aperture size and the quiescent radiation pattern. In the next section two categorically different methods of selecting the quiescent pattern to improve nulling resolution will be discussed.

### 3.2.2 Quiescent Pattern

The quiescent radiation pattern is usually selected to provide the desired coverage in a jam-free environment. This does not, in general, provide a best approximation to the desired coverage in the presence of a jammer. The adapted radiation pattern can be improved by either choosing the phase or the amplitude of the quiescent pattern appropriately. This will, in general, degrade the coverage provided in a jam-free environment.

In the next section the characteristics of a quiescent pattern with a linear-phase gradient and "uniform" amplitude will be calculated as a function of the magnitude of the gradient. This will be followed by calculating the radiation pattern when the quiescent pattern is enhanced in a direction near the jammer.

#### 3.2.2.1 Phase Distribution

Although several phase distributions could be considered, the linear-phase gradient has received the greatest attention. In an MBA it consists of exciting, or weighting, each beam with

an increasing, or decreasing, phase in the direction of desired nulling enhancement: the relative phase between adjacent beams is constant.

The foregoing can be presented quantitatively by considering an MBA with 5 beams centered on a line with equal angular separation $\theta_s$. Each beam is produced by uniform illumination of the same circular aperture with diameter D, and is represented by

$$E_n(\theta) = J_1(U)/U \tag{14}$$

where

$$U = (0.5\pi D/\lambda)\mathrm{Sin}(\theta - n\theta_s). \tag{15}$$

The antenna's radiation pattern is formed by weighting each beam and summing all of them at a single output port. The resulting antenna pattern is given by

$$E(\theta) = \sum_{n=1}^{N} A_n E_n(\theta) \tag{16}$$

where $A_n$ is a complex beam weight with magnitude between 0 and 1 and phase between 0 and 360 degrees.

For uniform coverage the beam spacing $\theta_s$ is chosen so that adjacent beams crossover at a point 6 dB below their maximum value (i.e., U = 1.8 in (14)). When adjacent beams are not excited in-phase (e.g., phase of $A_n$ not equal to the phase of $A_{n+1}$) uniform coverage is obtained by increasing the crossover level to a value greater than -6 dB. If the difference in phase between two adjacent beams exceeds 120 degrees it is impossible to achieve a uniform coverage pattern; the variation of the antenna pattern becomes severe when

adjacent beams are 180 degrees out of phase. In other words the crossover level between adjacent beams is determined by the angular spacing between them. This spacing can be adjusted to give uniform coverage as long as the magnitude of the phase of one beam, with respect to an adjacent beam, is less than 120 degrees.

Using (16) the radiation pattern of the MBA was calculated for a phase $\Delta\varphi$ between adjacent beams equal to 0, 45, 90, and 135 degrees. The results are shown in Figure 3-2 where the crossover level between adjacent beams equals -6 dB. Note the results indicate the ripple is about the same for $\Delta\varphi = 0$, 45, and 90 degrees. The ripple increases from about 2dB peak-to-peak to 10 dB peak-to-peak when $\Delta\varphi = 135$ degrees.

Nulling resolution of this ideal MBA is calculated by subtracting a maximum directivity beam (i.e., a beam with shape given by (14)), pointing in the direction of a jammer located at 0.5 degrees, from the pattern given by (16). Prior to subtraction, the maximum directivity beam is weighted so that a null is produced at $\theta = 0.5$ degrees. The results are shown in Figures 3-3a and 3-3b for $\Delta\varphi = 0$, 45, 90, and 135 degrees. The computed relative field was presented as -40.00 dB whenever it was less than -40 dB. These calculations were repeated with a -4-dB crossover level. The results are shown in Figures 3-4, 3-5, and 3-6.

Note the width of the null decreases with increasing value of $\Delta\varphi$. Referring to the -10-dB level, the null width decreases by a factor approximately equal to three if $\Delta\varphi$ is increased from 0 to 90 degrees. This is in agreement with results reported by Potts and Mayhan of MIT/Lincoln Laboratory [Reference 10]. This demonstrates that use of (14) and (16), as indicated, produces results that could be used to determine some of the design parameters such as $\Delta\varphi$ and the crossover level without the need for extensive calculation. Using these

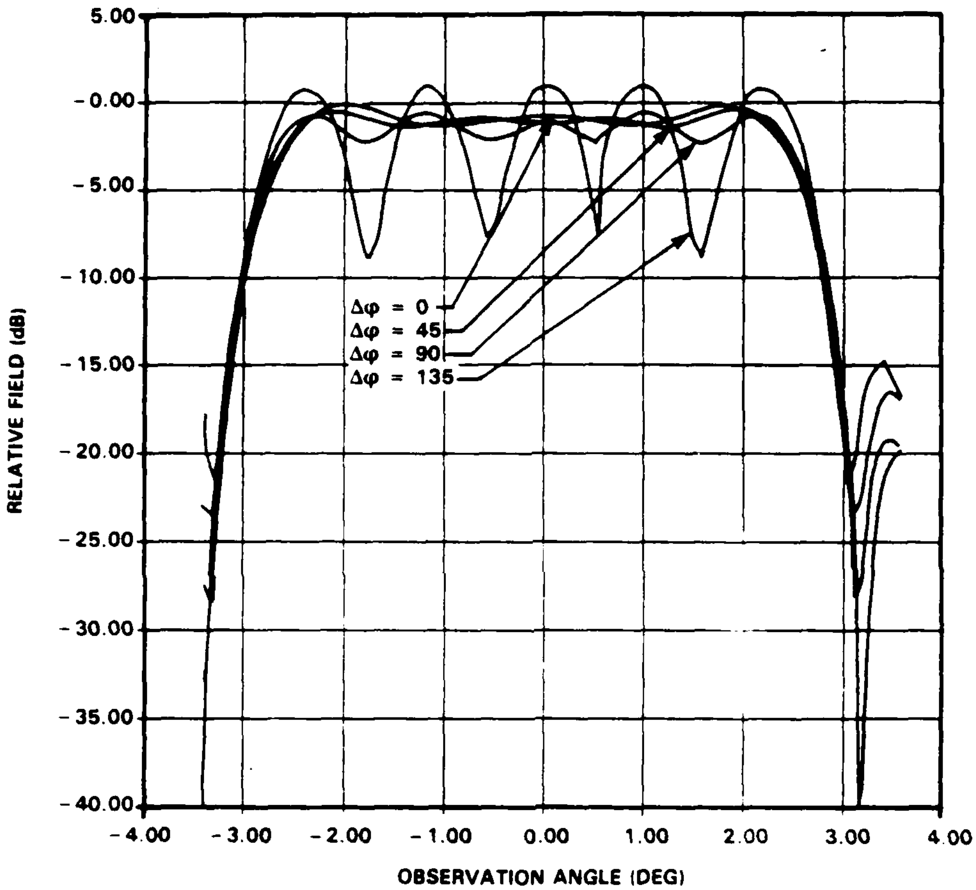



**Figure 3-2. (U) Phase-Tapered Quiescent Patterns: $-6$ dB Crossover, $D = 140\lambda$**

511015.0

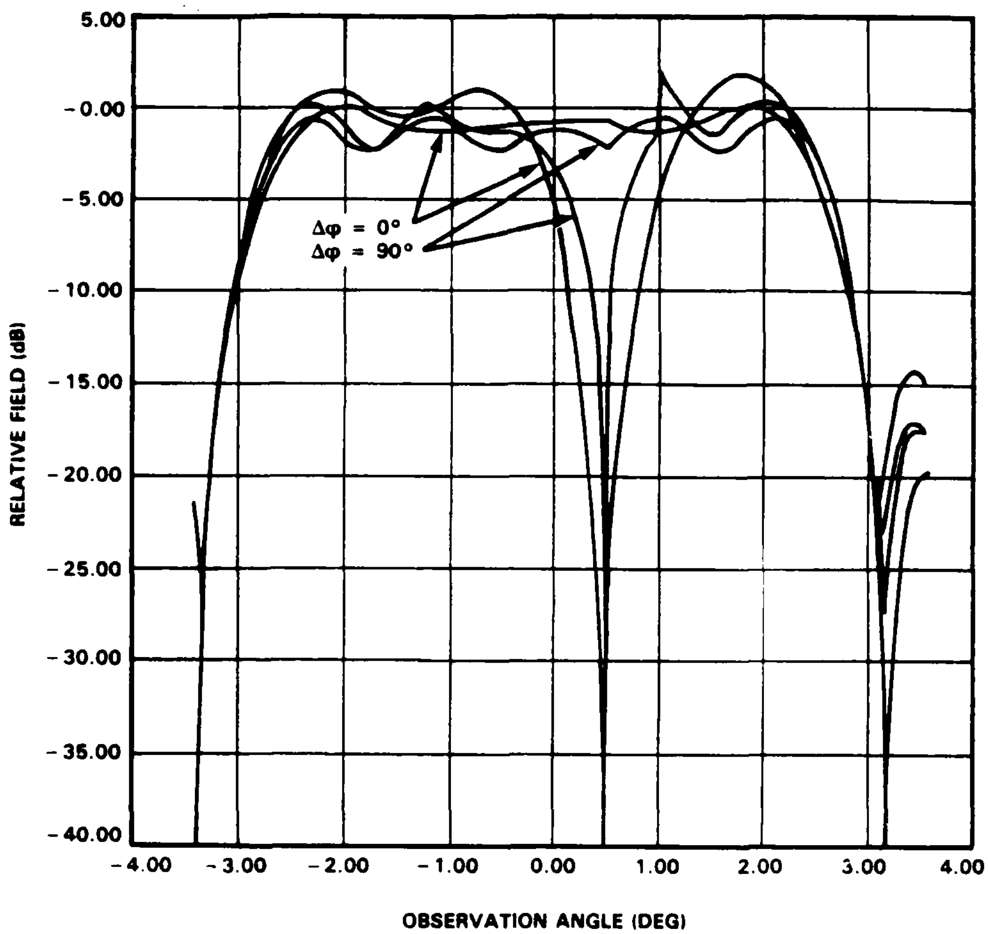
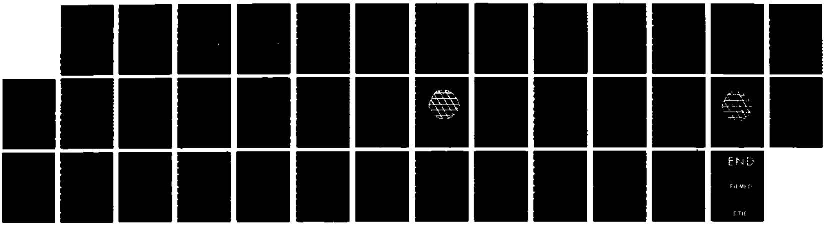
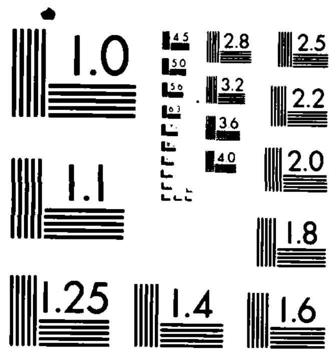



**Figure 3-3a. Adapted Patterns: − 6 dB Crossover with $\Delta\varphi$ = 0°, 90°, D = 140λ**

511005 1

AD-A162 116 DSCS (DEFENSE SATELLITE COMMUNICATIONS SYSTEM)/WIDEBAND 2/2
SHF (SUPER HIGH F (U) M/A-COM GOVERNMENT SYSTEMS DIV
VIENNA VA L RICARDI 29 NOV 85 DCA/MSO-85-3-PT-2
UNCLASSIFIED DCA100-84-C-0009 F/G 17/2 NL






MICROCOPY RESOLUTION TEST CHART
NATIONAL BUREAU OF STANDARDS-1963-A

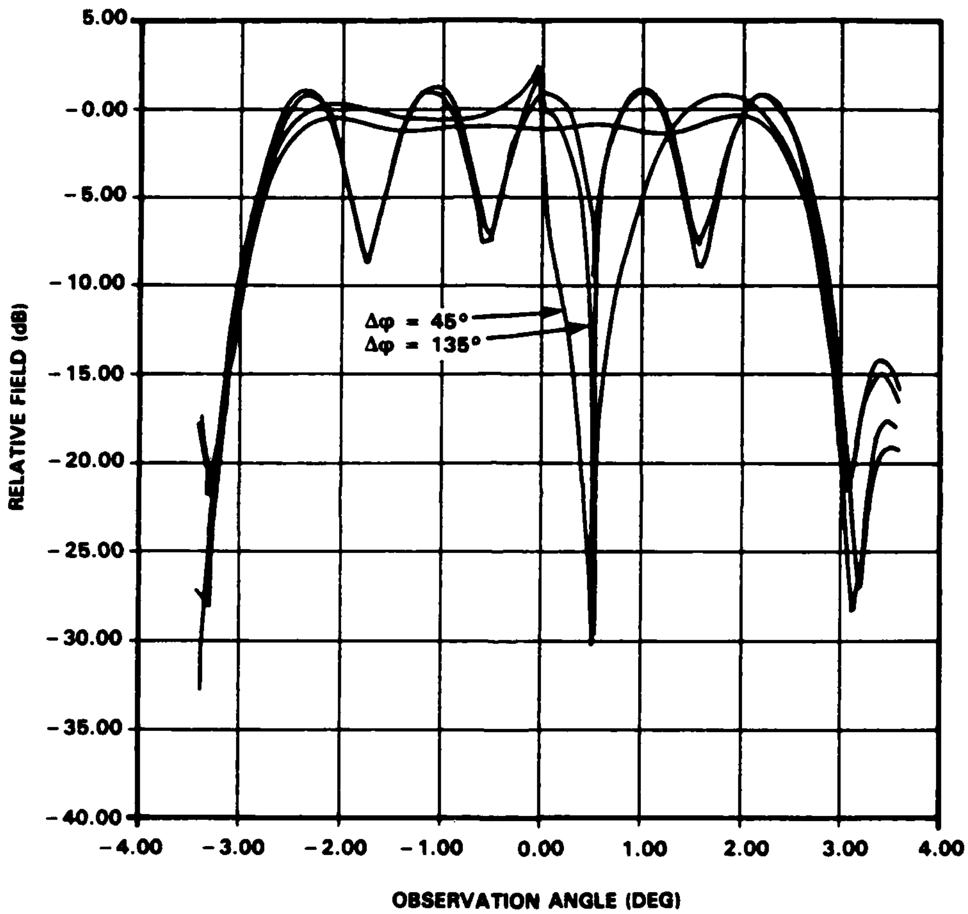



**Figure 3-3b. Phase-Tapered Adapted Patterns: −6 dB Crossover with $\Delta\varphi$ = 45°, 135°, D = 140λ**

511016.0

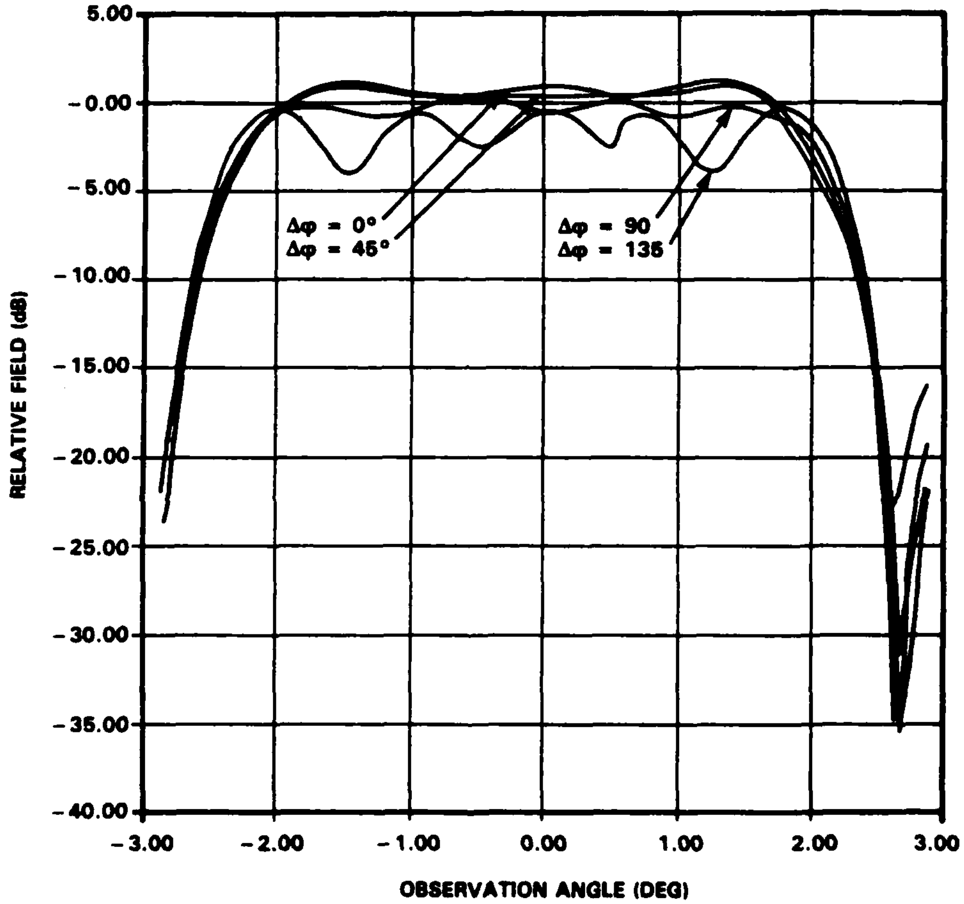



**Figure 3-4. Phase-Tapered Quiescent Patterns: −4 dB Crossover, D = 140λ**

511017.0

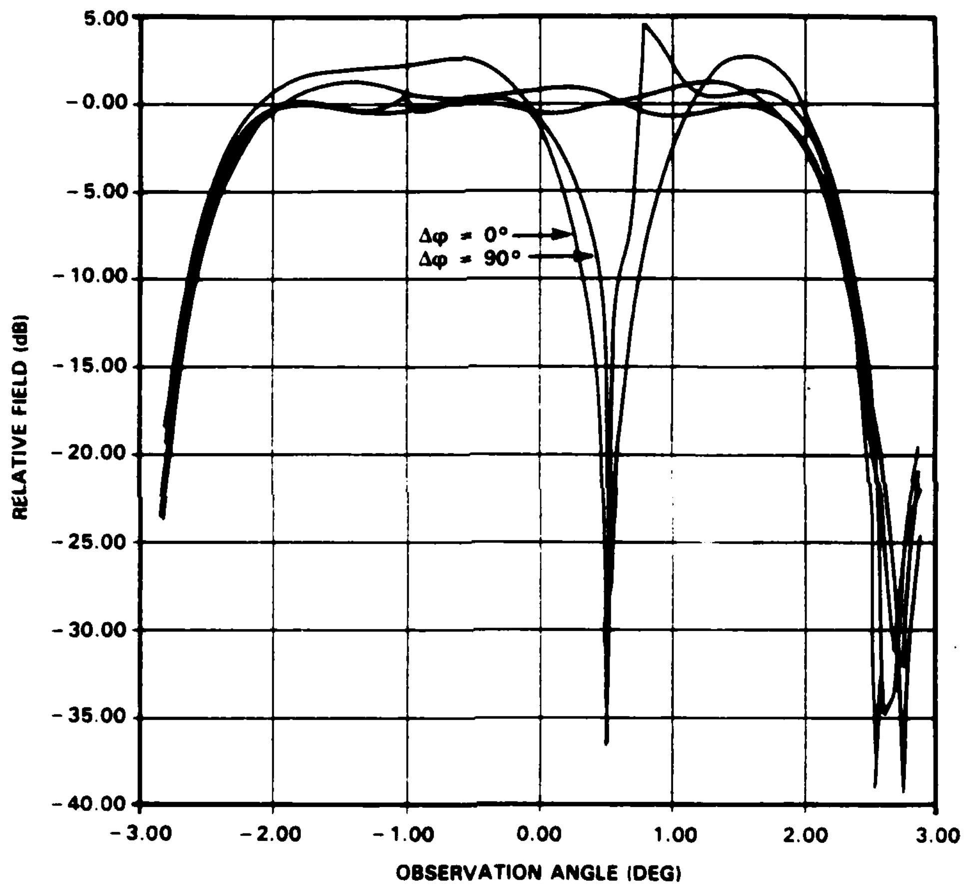



**Figure 3-5. Phase-Tapered Adapted Patterns: −4 dB Crossover with Δφ = 0°, 90°, D = 140λ**

511018.0

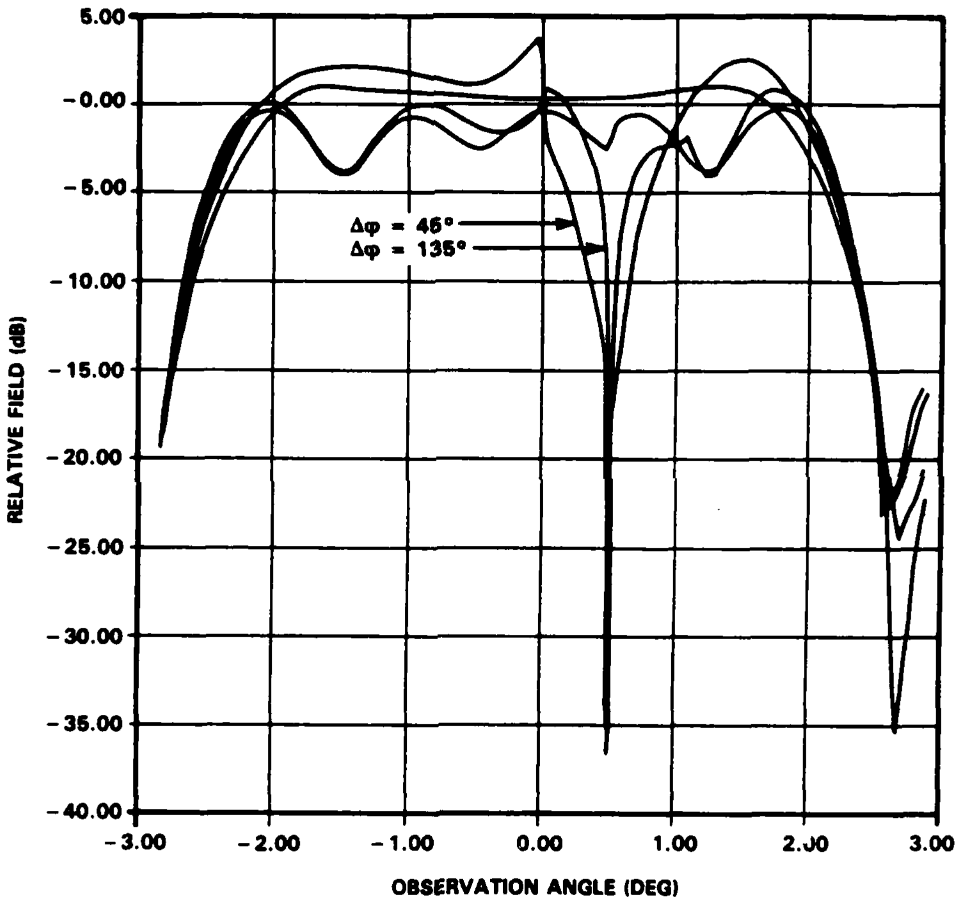



**Figure 3-6. Adapted Patterns: − 4 dB Crossover with φ = 45°, 135°, D = 140λ**

511019.0

parameters as a departure point, a more accurate representation of the MBA's beam patterns could be calculated and the parameters could be optimized.

Relative gain of the adapted and corresponding quiescent radiation patterns shown in Figures 3-3a and 3-3b were calculated by first recognizing that antenna gain is given by:

$$G = 4\pi / \int_0^{2\pi} \int_0^{0.5\pi} P \sin\theta \; d\theta \, d\varphi \qquad (17)$$

where $P = |f(\varphi)E(\theta)|^2$

and then assuming the pattern given by (16), with and without the subtraction of (14), is the same in the $\varphi$ plane regardless of the parameters that were varied to obtain the curves shown in Figures 3-3a, 3-3b, 3-4, and 3-5. Therefore the increase in gain, dG, when the null is installed, is given by:

$$dG = Ga/Gq, \qquad (18)$$

where Ga and Gq are computed using (17) and integrating only with respect to $\theta$.

### 3.2.2.2 Amplitude Distribution

Nulling resolution can be increased by increasing the gain (i.e., amplitude tapering) of the quiescent pattern in the direction of a user located near jammer (see Figure 3-7). This necessarily decreases the gain to essentially all other users in the coverage area; the decrease in gain depends on the increase in gain in the direction near the jammer.

Using the same MBA described by (16), this enhancement technique can be demonstrated by setting $A_3 = 1$ (i.e., the

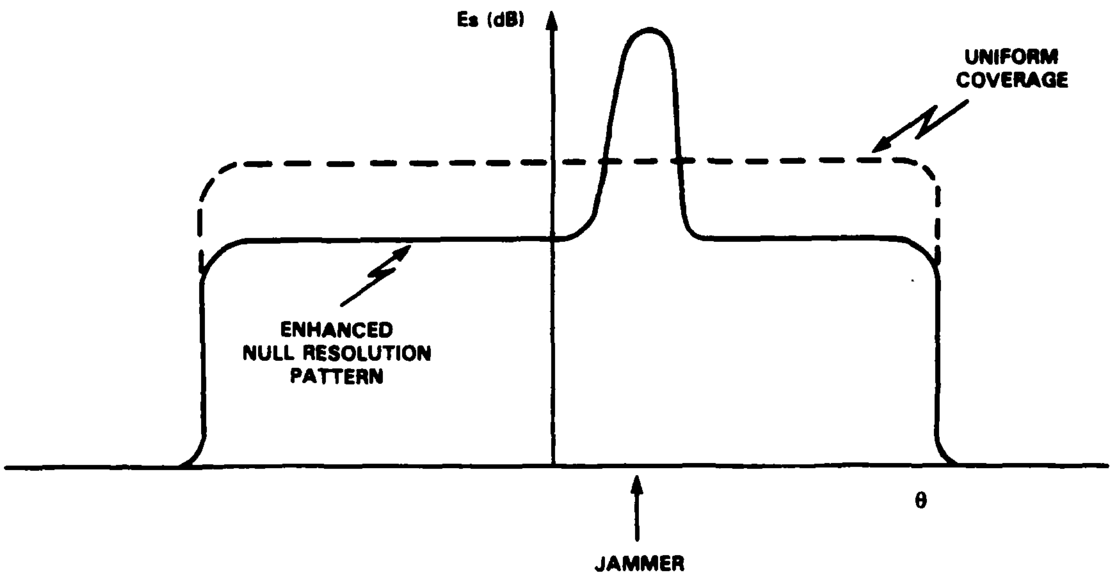



**Figure 3-7. Non-Uniform Quiescent Pattern**

511020.0

center beam amplitude = 1) and all other $A_n$ = .25. With the jammer located at $\theta$ = 0.5 degrees and $\Delta\varphi$ = 90 degrees, the quiescent and adapted patterns are as shown in Figure 1-7. The results obtained with $A_n$ = 1 are also shown for comparison.

Note the width of the null at -10-dB level decreased by a factor of about two when amplitude tapering is introduced. The resulting null width is less than 0.1 degree. This corresponds to a jammer-user separation less than 20 miles at synchronous altitude. This increased nulling resolution was obtained with an approximate 3-dB loss in gain in areas not near the jammer. Further increase in the gain toward the user, located near the jammer, would not result in a significant increase in the nulling resolution.

### 3.2.3 Sidelobe Canceller Consideration

Even the small exposure to the term "nulling resolution" given in the previous sections brings out the inherent difficulty to express performance in a universally agreed upon and understood value for the nulling resolution. Whatever the method of calculating the nulling resolution, increasing the aperture subtended by the antenna improves the nulling resolution. This is characteristically presented as the major advantage of a TPA over an MBA, or a filled array. In other words, a TPA with a larger aperture than an MBA can provide better spatial discrimination to a user located close to a jammer.

Unfortunately the TPA is much more vulnerable to jammers located a long distance from a user, perhaps in a sanctuary territory. On the other hand, the MBA is essentially invulnerable to a sanctuary-jamming scenario. It follows that a mix of the two configurations might result in the best of both. Toward this end the sidelobe canceller should be considered in conjunction with an MBA.

The sidelobe canceller gets its name from its initial use in suppressing the sidelobes of a radar antenna. It consists of a low-gain antenna (about 10-15 dB less than the gain of the radar antenna) supported at least a distance D/2 from the center of the radar antenna, where D is the diameter of the radar-antenna aperture, and pointed in the same direction as the radar-antenna beam. Signals received by the low-gain (auxiliary) antenna are weighted and summed with those signals received by the radar antenna. The weight is chosen to minimize the jamming signals that appear in the radar signal. Since the gain of the auxiliary antenna is less than the gain of the radar antenna, choosing the weight to minimize the total power received at the output of the summing circuit cannot significantly reduce the radar signal but it will place a null in the direction of the jammer.

This configuration was used in radar systems for convenience rather than increased resolution, even though it did improve nulling resolution. The same principle can be used to improve the resolution of a MILSATCOM uplink antenna when the latter is attempting to provide access to a single user located extremely close to a jammer. In particular, consider an MBA, with aperture diameter D, and three conventional center-fed paraboloids, with diameter D/3, equally spaced on a 3D diameter circle centered on the MBA's aperture. (See Figure 3-8). These auxiliary antennas will have a gain about 10 dB less than the maximum gain (i.e., with only one beam excited) of the MBA. Three antennas are used to ensure at least one antenna will be located so that a line joining its center with the center of the MBA, and the line joining the jammer and the nearby user, lie in the same, or nearly the same, plane.

This hybrid-nulling antenna configuration (Figure 3-8), uses a power-inversion nulling algorithm to set the weights of the auxiliary antennas prior to summing them with the signal received by the MBA. The algorithm will reduce the gain of the

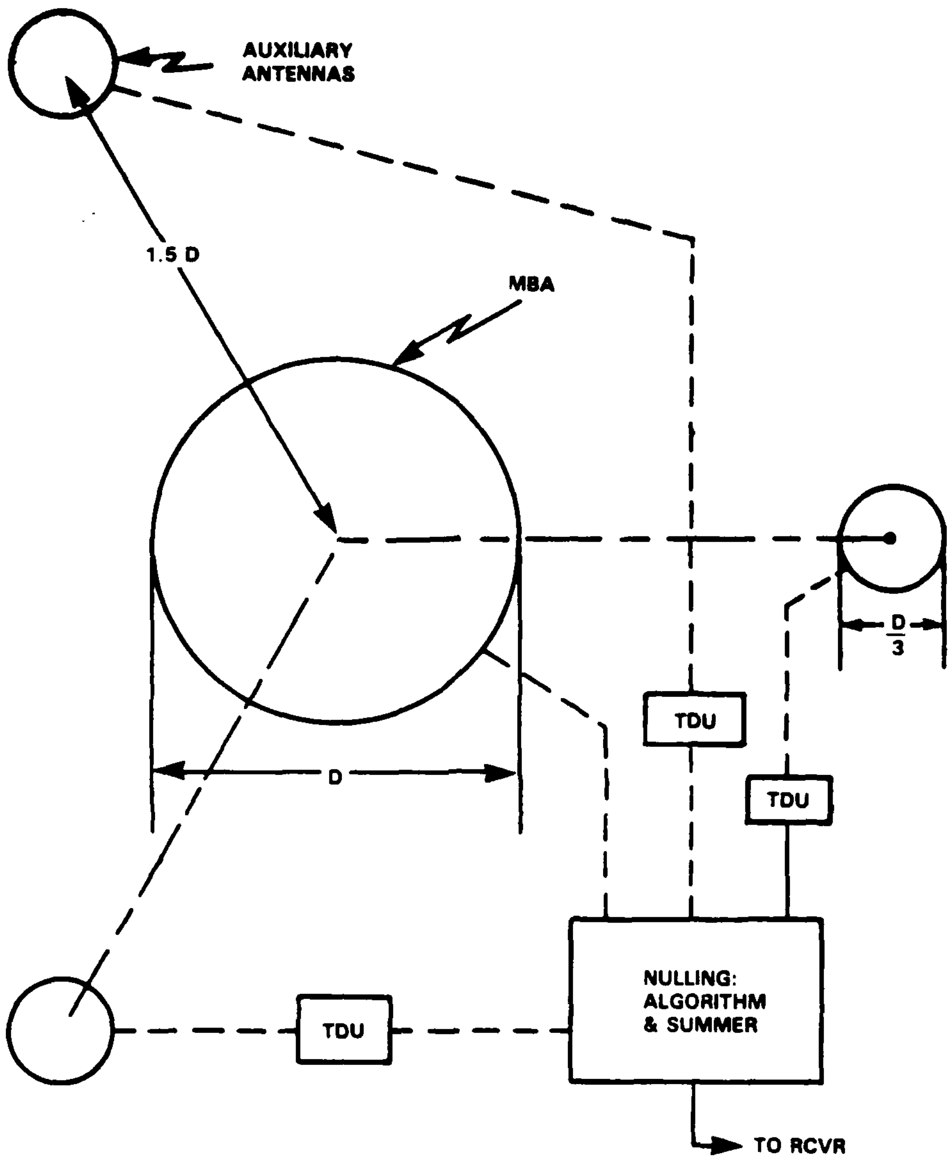



**Figure 3-8. Hybrid (Sidelobe Canceller) MBA Nulling Antenna**

511021.0

MBA if the jammer is located in a direction where the quiescent pattern gain is less than 5 dB below the peak of the MBA's beam; otherwise it will reduce the gain of the auxiliary antennas. It is assumed the auxiliary antennas are pointed toward the jammer. The approximate radiation pattern is given by:

$$E_s(\theta) = J_1(u)/u - A_s(\exp(j(\theta_w - \psi(1-2\cos 60))) \quad (18)$$

where

$$\psi = 3\pi D \sin(\theta)/\lambda$$

and $A_s \exp(j\theta_w)$ is chosen to reduce $E_s$ to 0 in the direction of the jammer (i.e., $\theta = \theta_s$).

Using (18) $E_s$ was calculated for $D = 140\lambda$ as in the calculation of the data for Figures 3-2, 3-3a, 3-3b, and 3-4. The jammer was located at $\theta j = 0.5$ degrees. The potential for the improved nulling resolution through the use of auxiliary antennas can be obtained by comparing the results shown in Figures 3-2, 3-3a, 3-3b, and 3-4 with those in Figure 3-9.

As it is described here, the hybrid system (sidelobe canceller with an MBA) has several disadvantages, namely:

1. The auxiliary antennas must be pointed toward the jammer because their radiations patterns are not wide enough to cover the earth FOV.

2. The potentially large differential time delay between signals received on the four antennas will reduce the band width over which the null is maintained.

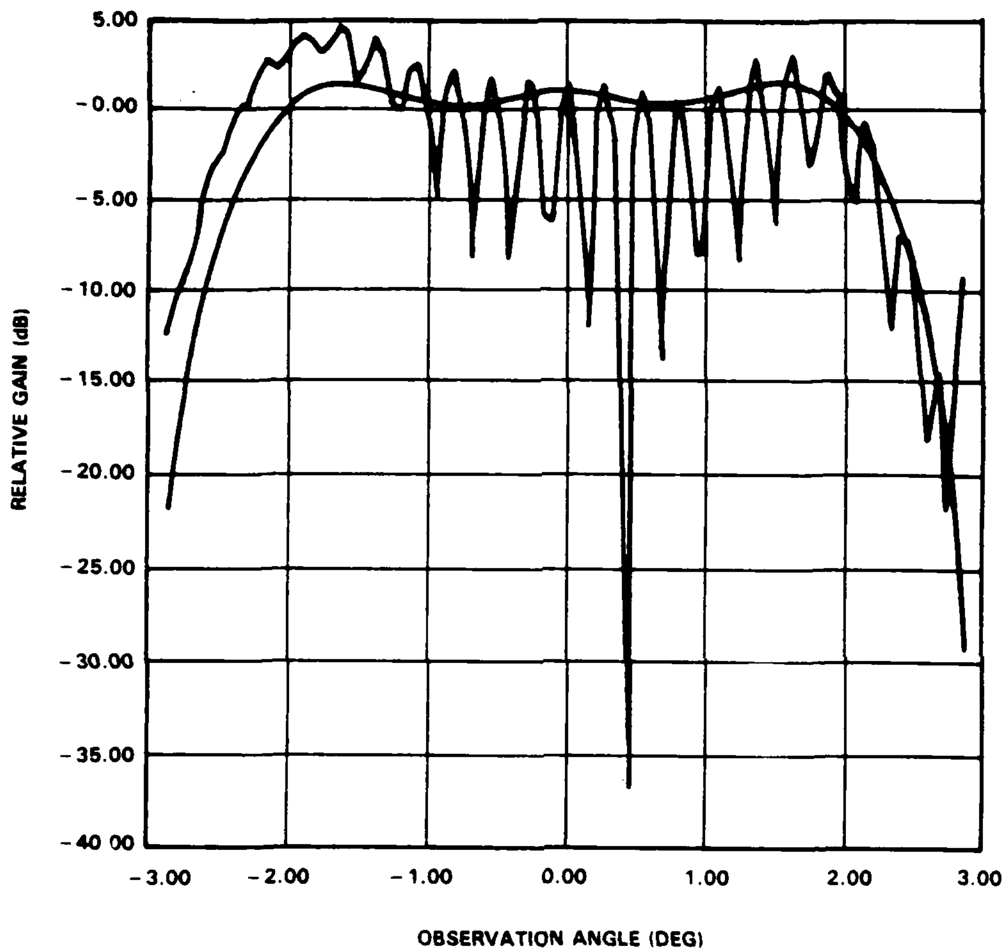



**Figure 3-9. Hybrid MBA/Phased-Array Antenna Pattern (Adapted vs Unadapted −4 dB Crossover)**

511022.0

3. Three additional antennas and their associated circuitry are required.

4. A substantial amount of analysis is required to determine the performance of the configuration.

This and the discussion in Section 3.3 are presented to suggest further consideration of this configuration with appropriate analysis.

### 3.2.4 Threat Considerations and Tradeoffs

User requirements and threat definitions are generated initially without knowledge of the physical realizability of meeting the requirements and satisfying the threat scenarios with a given implementation. It should follow that requirements and threat definitions that drive the system design to an unreasonably expensive or risky implementation must be revisited and benefit from the knowledge gained during a first iteration. The current ESAAP requirement that a jammer can be located very close to a user in a tactical battle area requires an antenna aperture too large to implement in any configuration other than the TPA. However, this design must use earth-coverage antennas as elements in order to be economically and physically acceptable. Use of earth-coverage elements renders the antenna vulnerable to distant jammers in sufficient numbers to disable the communication system.

Increasing the EIRP of terminals located near a jammer will improve the antijam capability of the terminal proportionately. This increase can be obtained by increasing the antenna aperture, the power output of the HPA, or both.

Specifying the jammer is in the clear and the user is in rain is certainly a worst-case scenario but is not a realistic scenario. The probability that an enemy will deploy a jammer

that depends on that occurrence is highly unlikely because it is indeed a rare event. It will occur substantially less than two percent of the time. The economic investment required to build a jammer that will be effective less than two percent of the time would be difficult to defend. The enemy would only invest funds to build a jammer that would not be dependent on such a favorable scenario or find some other method to disable communication, command, and control.

## 3.3 CANDIDATE ANTENNA

Over the past few years several antennas have been suggested as a candidate for an EHF MILSATCOM system and many have been studied in sufficient depth to obtain a reasonably good understanding of their relative merits, performance capability, risk, and maturity. It is not the intention of this report to describe a new EHF uplink antenna that will meet all the known user requirements. Rather it is intended to present a "best" candidate antenna system as a baseline, to support previous claims to its expected performance capability, and to consider modification of it to meet more of the user requirements.

The better of several candidate systems were reviewed at the DSCS III Upgrade Working Group Meeting held at Defense Communications Engineering Center (DCEC) on April 2-3, 1985 [Reference 11]. At this meeting Dr. Joseph Mayhan, of MIT-Lincoln Laboratory, described an MBA configuration he said performed satisfactorily in all the scenarios considered. Mr. Tom Treadway described TPA configurations that were modifications of previously studied antenna systems. He also stated that the antenna performance was adequate for the scenarios considered. However it was generally agreed the TPA is vulnerable to sanctuary jammers that cannot affect the performance of the MBA. This vulnerability of any TPA configuration should be recognized as a major deficiency

suggesting that the MBA configuration is the better candidate; this is especially true if the MBA can be augmented to provide the required suppression of near-in jammers where the TPA typically excels. However, vulnerability to near-in jammers is not a major deficiency of an MBA since only a few, perhaps one, terminals are disrupted and the jammer might be removed physically, if necessary.

Using the foregoing tools, MBA configuration will be considered and the expected performance capability will be discussed. This performance will be shown to agree with that given by Dr. Mayhan. With the addition of a sidelobe canceller, it will be shown near-in jammers can be suppressed while still maintaining adequate margin to nearby users. It is necessary to first place various link budgets in evidence since they will provide the parametric relationships necessary to predict expected performance capability. A parametric relationship is presented to avoid restricting the use characteristics of extant, or planned, terminals.

The link budgets shown in Table 1-1 are for ground terminals with antenna apertures, transmitter output power, and EIRP ranging from two feet, 25 watts and 58.8 dBW to 40 feet, 5 kW and 108 dBW. All terminals operate at 44.5 GHz and use data rates from 100 kbps to 100 Mbps; the data rate indicated for each terminal is suggested as a strawman value. The indicated data rate can be increased, or decreased, with a corresponding change in the indicated margin. The satellite antenna is assumed to have a 24-inch diameter aperture and a combined pointing and RF circuit loss equal to 3.5 dB. Notice the link margin varies from 6.9 dB to 26.2 dB when the terminal is operating in a rain storm and undergoing 12 dB of rain-induced attenuation in addition to the assumed 2.34-dB atmospheric (no rain) attenuation. In other words, the indicated margins are 12 dB larger when the terminal is operating in a rain-free environment.

It is important to note that over the earth's disk, rain-induced attenuation at EHF exceeds 12 dB less than two percent of the time. In many locations rain attenuation exceeds 12 dB less than .1 percent of the time. It is also true that rain attenuation can exceed 20 dB during rare but significant periods of time. In view of this it is not necessarily wise to design the satellite communication (SATCOM) system to overcome large-rain attenuation by increasing the terminal's EIRP or the satellite's G/T. Rather it is wiser to use three or more terminals operating cooperatively in a space-diversity mode or to accept the expected outage or reduction in data rate during periods of intense rainfall. It may also be wise to trade rain margin for near-in jammer suppression in those scenarios that disrupt communications of users located near a jammer. (This is principally a performance evaluation criteria rather than a design goal). That is, a terminal can be jammed 100 percent of the time whereas it may not be able for the jammer to disrupt communication if the rain margin is used to increase spatial discrimination. If the rain margin is used for this purpose, the jammer is effective only when it is raining. It is very unlikely an enemy would build a jammer that could operate successfully only when it is raining.

### 3.3.1 Description

Any upgrade of DSCS III to EHF must meet, or satisfy, the needs of the current and projected communities of users. If a proposed design meets the desired performance characteristics but exceeds available launch weight, power, etc., it is necessary to modify the design and/or the desired performance characteristics. Over the past 5 years, several payload configurations have been considered, covering a wide range of capability. High data rates and high resolution spatial discrimination have been the major drivers in the creation of these configurations. The spatial discrimination requirement has given rise to two basic antenna configurations: a TPA and

a MBA. It is generally agreed that, for a given communication capacity and satellite weight and power, the TPA has superior close-in jammer discrimination and the MBA's performance is superior for all but the close-in jammer scenarios. It is also true that a system using the TPA can be completely disabled by a few more than N jammers (N = number of elements in the TPA) located, thousands of miles from a user in a protected territory; whereas, a system using a MBA is invulnerable to jammers outside of the user's instantaneous coverage area. It is also generally agreed that adaptive-nulling MBA systems are much more mature than TPAs. Extant systems such as DSCS III and the MILSTAR nulling antenna support but do not prove this latter statement because significant research and development and test are currently being planned for the ESAAP phased-array antenna.

In this section a candidate antenna configuration is described. Its performance limitations are estimated, using tools developed in the previous section, and its weight and power are estimated using data gathered by MIT-Lincoln Laboratory. The proposed payload falls slightly short of meeting some of the requirements; increasing antenna size and weight could result in a payload meeting all requirements except it would have excessive weight and size. A modification to the basic system is proposed to perhaps result in a payload that satisfies all requirements including size and weight. The modification results in a hybrid antenna with the fundamental performance characteristics of both the MBA and the TPA with about the same weight and power.

#### 3.3.1.1 Earth FOV

In the current context, earth FOV means a timeshared basis as opposed to instantaneous. Specifically one, or more, uplink, high-gain beams are TDMA switched to offer access to users located anywhere on the earth's disk. At any instant a

single user accesses the satellite (i.e., on the uplink) via a beam; with B beams simultaneously in operation, B users could access the satellite at the same time. The links set up in this manner are often referred to as "point-to-point" communication links. The downlink antenna would probably be an active aperture antenna [see Reference 9] or a gimbal-mounted paraboloid depending on the required downlink capacity.

The coverage capability area assumes several simultaneous users in a beam broader than a single beam of the MBA and operating in a FDMA mode, perhaps in a TDMA mode, with similar communities of users. This division of users is driven by their data rates and the satellite's communication capacity. In other words, it is important for users to operate either in a continuous-wave (CW) mode or at a greater than 10 percent duty cycle if they operate in a TDMA mode. (This is discussed later in Section 3.4.2). Low-data-rate users should operate together in a FDMA mode. If there is insufficient bandwidth to handle the entire low-data-rate community, it will be necessary to either use demand assigned multiple access (DAMA), which is a form of very slow TDMA, or FDMA/TDMA with subsets of the low-data-rate users sharing the allocated satellite facility in a TDMA mode. Medium-data-rate users may fall into either the area coverage or point-to-point categories depending on the total data rate of the community and the bandwidth allocated to them as a group.

The foregoing multiple access discussion will be examined in greater detail in the following. Assume there are three classes of users 1) high-data-rate users who operate at data rates greater than 2 Mbps; 2) medium-data-rate users who operate at data rates between 100 kbps and 2 Mbps; and 3) low-data-rate users who operate at data rates less than 100 kbps. It is further assumed the number of low-data-rate users is at least ten times greater than the number of medium-data-rate users and they in turn outnumber the high-data-rate users by a

factor of about ten. Let us further assume that the combined data rate of all simultaneous users is less than or equal to the throughput capacity of the spacecraft's payload.

It is probable the high-data-rate users will require maximum satellite G/T in order to support their high data rate. Consequently, they will be provided access through a maximum directivity beam with perhaps less than 10 high-data-rate users accessing the satellite at the same time using TDMA. In sharp contrast the low-data-rate users will require a relatively large area coverage beam and consequently lower G/T than for the high-data-rate users. The lower G/T is consistent with the low data rate and lower terminal EIRP of the low-data-rate users. However, there may be more than 10 low-data-rate users wishing to access the satellite simultaneously; consequently, they will have to operate in a FDMA/TDMA mode. That is the satellite's area coverage beam will be switched among the low-data-rate user communities providing each community with timeshared FDMA. The medium-data-rate users will also operate with TDMA and in some cases FDMA/TDMA with a few users operating in a high-gain uplink beam.

This segregation between users of different data rates leads to a baseline requirement for the satellite to generate at least one switched uplink beam per community of users. The recommended EHF uplink antenna is a MBA consisting of a 24-inch diameter lens illuminated by an array of 285 feed horns whose ports are connected to a 16 degrees-of-freedom (DOF) adaptive algorithm via a 285:16 switching tree using ferrite-isolator latching switches. This switch system is made up of 16 switch trees; each tree has up to 19 inputs any one of which can be connected to a single output port.

The overall antenna system is sketched in Figure 1-5. The beam number and location is presented in Figure 3-10 where the array of feed horns is represented by 19 groups of feed horns.

Each group of feed horns are numbered 1 through 16. (Some groups have less than 16 feed horns). All number 1 feed horns are connected to a single input (i.e., the number 1 input) to the 16 DOF nulling algorithm via the number 1 switching tree. Similarly the number 2 feed horns are connected to the number 2 input to the 16 DOF nulling antenna via the number 2 switching tree, etc.

This is not a new MBA concept, it is very nearly identical to the one presented by Dr. Joseph Mayhan at the Working Group held in early May 1985 at DSCS [Ref. 11]. It is also similar to one described in Reference 12. It is important to point out that each feed produces a beam with a HPBW approximately equal to 0.7 degrees and the beam centers are arranged on a triangular grid with contiguous coverage equal to the earth's disk as seen by a synchronous satellite. Triangular spacing improves the spatial resolution and permits smaller jammer-user separation than other beam arrangements. Feasibility of this design has been demonstrated by analysis, partial breadboard assemblies, and models conducted and/or prepared by MIT-Lincoln Laboratory.

3.3.1.1.1 MBA

The proposed switch tree can select any 16 of the 285 beams (that is, one number 1, one number 2, etc., over number 16) and connect them to the nulling algorithm in less than 1 microsecond. By ground control the users are allocated a time slot and a cluster of four, or more, adjacent beams point toward the user terminal at the appropriate time during a frame. During each frame, a beam points toward all users, currently accessing the satellite, at least once. The cluster of four, or more, beams is required to suppress local jammers; the quiescent pattern is chosen to provide maximum gain toward the user. Phase and shape of the quiescent pattern is chosen

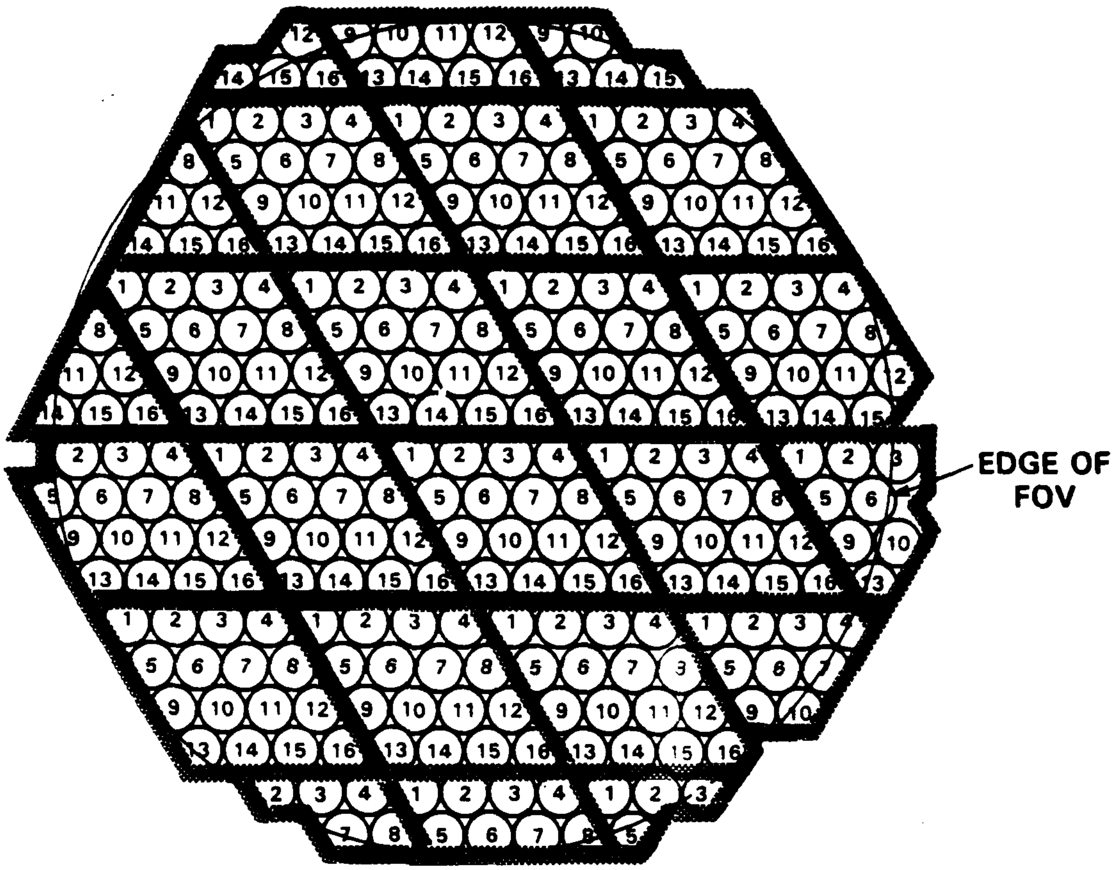



Figure 3-10. Beam/Feed Horn Number and Location

to improve the nulling resolution. These pattern characteristics can be obtained using ground control and a knowledge of the jamming scenario, or by on-board processing (see Section 3.3.1.2.1).

In principle, up to four simultaneous beams can be produced to serve four different users or user communities. However, the possibility of more than one user needing, say, a number 1 beam leads to a conflict. This conflict might be mitigated by appropriate time slot allocation, or through a DAMA discipline. Only a traffic analysis will yield an accurate understanding of this potential conflict.

Referring to Table 1-1, a terminal with a 40-foot-diameter antenna aperture and transmitter output power equal to 5 Kw could communicate at 100 Mbps with a 26-dB margin. Reducing the antenna diameter to 30 feet and the transmitter's output to 2 kW and increasing the data rate for 1 Gbps reduces the margin to 9 dB. Because of these high data rates, antijam processing gain through spread spectrum is less than 10 dB; however, it may still be worth the implementation cost.

Spatial discrimination can be estimated using Figures 3-3a, 3-3b, or 3-5. Using Figure 3-3a notice the -9-dB width of the null is about 0.6 degree for a quiescent pattern with uniform phase and amplitude. The -9-dB null width decreases to about 0.2 degrees if the phase between adjacent beams equals 90 degrees (i.e., $\Delta\varphi$ = 90 degrees). The null automatically points toward the jammer suppressing his signals by more than 20 dB. In doing so, the gain to a user located 0.1 degree from the jammer will experience a 9-dB decrease in the satellite's uplink G/T reducing its margin by 9 dB. However, the curves in Figures 3-3a, 3-3b, etc., are for a 140$\lambda$-diameter antenna aperture; the proposed 2-ft aperture is about two-thirds this size. Therefore, the proposed 2-ft-diameter satellite antenna suppresses a jammer and simultaneously supports a terminal,

with EIRP = 100 dBW, operating at 1 Gbps when it is 60 miles from a nearby jammer that is in turn located near the satellite's nadir. This minimum separation increases to about 180 miles when the terminal and the jammer view the satellite at a 20-degree elevation angle; i.e., at the edge of the earth-coverage area.

Decreasing the terminal's EIRP decreases the data proportionately. For example, the lowest power terminal in Table 1-1 has 43 dB less EIRP than the one in the second column from the left. Consequently, it could operate in the same jammer scenario at about 50 kbps and no margin, but it would have to use some spread-spectrum antijam protection. Note this small terminal has a 2-ft-diameter antenna aperture and a 25-W high-power amplifier (HPA). Jammers at greater distances will also be suppressed either by the nulling algorithm or because the MBA rejects all signals arriving from those directions and corresponding beams that are not "connected" to the nulling algorithm (i.e., they have been switched off).

As the quiescent pattern is shaped to produce more than one uplink beam, the satellite G/T decreases. If the beams are separated more than a few beamwidths, G/T decreases as $1/N_b$ where $N_b$ is the number of beams turned on at a given time. If the beams are adjacent to one another the G/T decreases approximately as $1.6/N_b$. That is, if all 16 beams are weighted to produce a 2.8-degree-by-2.8-degree coverage area, the satellite's uplink G/T will decrease about 10 dB, not 12 dB. It may be possible to reduce the minimum tolerable jammer-user separation further by both amplitude and phase weighting the quiescent pattern. Study of the expected performance is beyond the scope of this report.

Although a lens antenna is indicated in Figure 1-5 an offset reflector with Cassegrain feed may give satisfactory performance and be somewhat lighter and less vulnerable to environmental conditions.

### 3.3.1.1.2 MBA and Sidelobe Canceller

Adding gimbal-mounted dishes as indicated in Figure 3-8 can increase the nulling resolution of the MBA (uniform quiescent pattern) by a factor of three or more. Unfortunately a significant increase in hardware is required and implementation problems not yet solved by the TPA studies remain to be solved. These problems include:

1. Implementation of a variable time delay unit.

2. An algorithm for setting the variable time delay unit.

3. Vulnerability to jammers outside of the MBA's instantaneous FOV.

4. Bandwidth over which a null can be maintained.

Nevertheless the simple analysis presented in Section 3.2.3 indicates significant improvement in nulling resolution.

### 3.3.1.2 Area Coverage

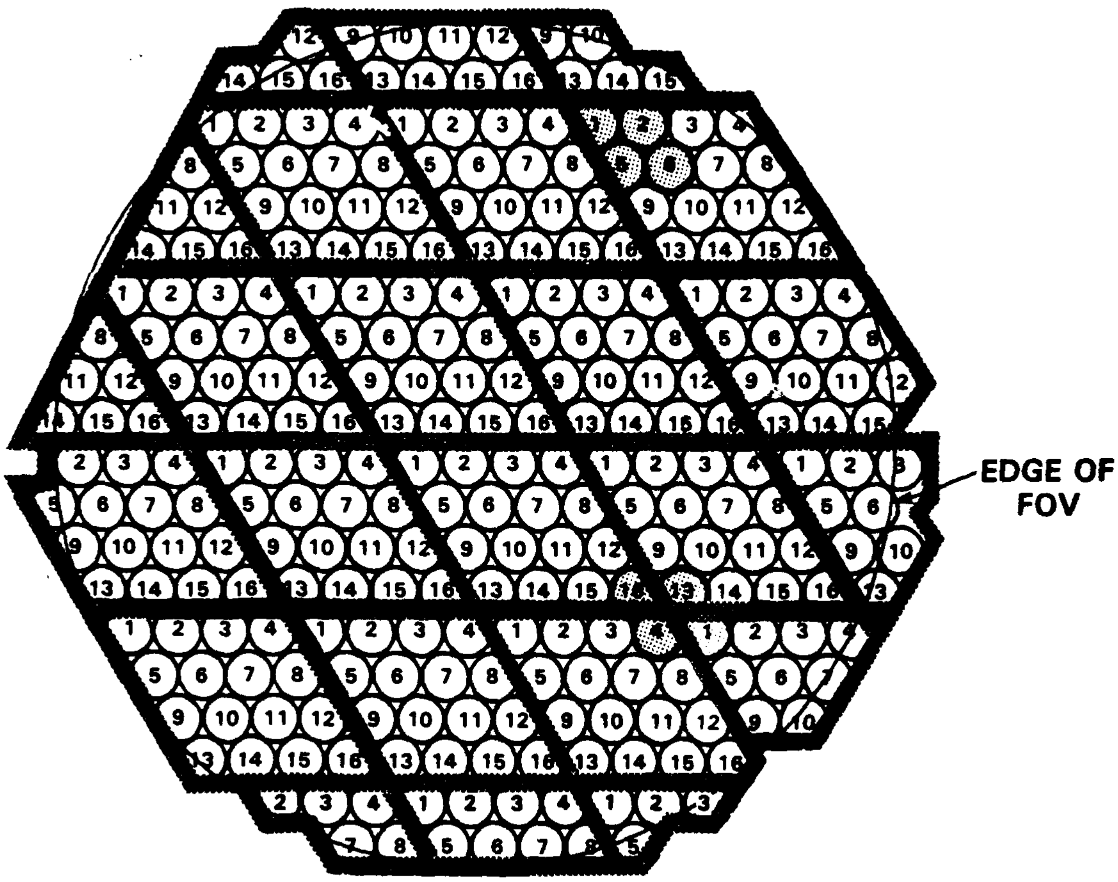
When the user community occupies an area subtending more than 0.7 degree, measured at the satellite, it is necessary to shape the quiescent pattern accordingly. This results in a decrease in gain of the satellite's uplink antenna. Assuming the coverage area will be less than 1.4 degrees square, the "pointing loss" introduced by weighting four adjacent beams equally (i.e., to provide the desired coverage) will be about 4.5 db. The margins given in Table 1-1 would decrease by the same amount. It is important to note the proposed antenna system can provide an area coverage pattern by setting the quiescent weight appropriately. Multiple beams can be realized in the same manner.

### 3.3.1.2.1 MBA

The candidate antenna functions essentially in the same manner as describea in Section 3.3.1.1.1 and requires no additional hardware to produce an area-coverage beam. The quiescent pattern is generated by setting the quiescent weights to effectively "turn on" those beams that serve the desired coverage area. These beams must first be selected by the 16 switch trees. The controller for carrying out this function must have knowledge of the user location and perhaps a general knowledge of the jammer location. The latter may be required to install the appropriate phase taper or amplitude "taper."

Alternatively, an on-board controller could "try" various phase gradients or pattern shapes and determine the best set of weights by trying, as with the power inversion algorithm, to minimize the power at the output of the summing circuit. This control is indicated functionally in Figure 1-5.

Creation of several simultaneous beams can require the use of the same number beam in two different simultaneous beam-coverage areas. This is indicated pictorially in Figure 3-11 where two coverage beams are desired simultaneously; one requires beam numbers 1, 2, 5, and 6 in the upper right quadrant of the MBA's FOV. The other requires beam number 1, 4, 13, and 16 in the lower right quadrant of the MBA's FOV. These horns are indicated by shading. In this scenario beam number 1 is in conflict. This is especially disadvantageous if the number 1 beam is absolutely essential to both coverage areas (i.e., it is used to modify the desired pattern). This conflict might be resolved by assigning different time slots to these coverage areas in the TDMA timeframe.




511024.0

**Figure 3-11. Beam #1 Conflict Scenario**

3.3.1.2.2. MBA and Sidelobe Canceller

It was pointed out the sidelobe canceller must use antennas with gain greater than -10 dB with respect to the MBA. Otherwise, the MBA's adapted pattern gain may be decreased more than a tolerable amount when the null is formed on the jammer. When operating in the area-coverage mode, the MBA's gain may be reduced as much as -9 dB (i.e., all 16 beams are "on" to produce a 2.8 degree-by-2.8 degree coverage pattern) and referred to its gain when a single beam is "turned on." This reduction in gain may permit use of small auxiliary antennas as sidelobe cancellers. Details of a hybrid, or sidelobe canceller, system should first be evaluated for that type of scenario requiring the highest nulling resolution.

### 3.3.2 Nulling Resolution

Results shown in Figures 3-3a, 3-3b, 3-5, and 3-6 substantiate claims that shaping and phasing the quiescent radiation pattern improves the nulling resolution by a factor of three or more. Assuming the nearby user can suffer a 9-dB decrease in the satellite antenna's gain when the adapted pattern is installed, the minimum tolerable angular separation $\theta_{ju}$ between the jammer and the user is approximately equal to 70 divided by the antenna aperture in wavelengths (i.e., a HPBW). Using either a phase gradient (i.e., about 90 degrees per beam spacing), or amplitude shaping in the quiescent pattern reduces $\theta_{ju}$ to about one-third of a HPBW. Notice that increasing the antenna's aperture (i.e., as in a TPA or a sidelobe canceller) by a factor of three results in the same improvement in nulling resolution.

### 3.3.3 Null Depth and Bandwidth

Studies at MIT-Lincoln Laboratory and TRW indicate null depths greater than 30 dB can be realized over a four percent frequency bandwidth. The proposed antenna would use a nulling algorithm similar to those used by MIT-Lincoln Laboratory and TRW. The switch trees are similar to those that will be used on MILSTAR; data on these devices indicates the candidate antenna system will achieve about the same null depth over the operating frequency band.

## 3.4 PROCESSING ISSUES

The TDMA-switched beam and nulling characteristics of the candidate antenna interact with the mode of multiple access used, the adaption time, and the tolerable dynamic range of jammer signals. Interaction among these characteristics is discussed in this section together with familiar signal processing considerations.

### 3.4.1 Adaption Time and Dynamic Range

The time required to set the receiving antenna weights to produce an adapted pattern (i.e., a pattern with the desired coverage and a null in the direction of all jammers) depends on the specific scenario. The expected maximum-and-minimum time are a function of the ratio of maximum to minimum jammer signals at the input to the nulling processor. With a single jammer present, adaption time is inversely proportional to the magnitude of the received jammer signal and the gain $G_f$ of the adaptive nulling processor's feedback loop. Designers usually set $G_f$ just large enough for the nulling loop to respond to the expected weakest jamming-signal level. Stable operation of the nulling loop is maintained by designing the nulling loop so that the largest expected jammer-signal level will not produce unstable operation. This desired dynamic

range is usually about 30 dB, and this results in a ratio of maximum-to-minimum adaption time approximately equal to 1000. A larger dynamic range tends to drive the receiver design; perhaps making it impossible or at least very difficult.

This limitation implies two constraints:

1. Maximum expected jammer-signal levels should determine $G_f$ the gain of the nulling feedback loop.

2. Jammer signals, less than -30 dB with respect to the expected maximum jammer signal, will probably not be sensed and nulled by the nulling algorithm.

The first constraint is self explanatory and straightforward. The second constraint implies that weak jammers are more likely to disrupt a communication system than are the stronger jammers. Using a variable $G_f$ may remove the systems vulnerability to weak jammers if the maximum expected jammer is not present. Use of spread-spectrum antijam, or inherent spatial discrimination, may also increase the tolerable jammer EIRP and the minimum jammer signals to which the adaptive nuller must respond.

### 3.4.2 Antijam Characteristics of TDMA vs FDMA

Fundamental multiple access modes of operation suitable for SATCOM systems usually are limited to TDMA, FDMA, and code division multiple access (CDMA). Of these, CDMA does not interact with the antenna design of either the terminal or the spacecraft. Consequently only the antijam characteristics of TDMA and FDMA will be discussed in this section.

### 3.4.2.1 Peak Power Limited

Terminals operating in a TDMA mode timeshare the satellites uplink. This is usually carried on a repetitive basis with each terminal in the TDMA net taking its turn to transmit once during a TDMA frame. The ratio of time a terminal transmits during a frame to the time duration of a frame is called the terminal's duty cycle $T_c$. If a terminal's average data rate is R, this terminal must transmit at a data rate $R_t$ when it is transmitting. Note:

$$R_t = R/T_c. \tag{19}$$

If $T_c = 1$ the terminal transmits continuously and the transmitted energy per bit is

$$E_b = P_t/R. \tag{20}$$

where $P_t$ is the CW, or average, power transmitted.

When operating in a TDMA mode (i.e., $T_c$ less than 1) the transmitted energy per bit is

$$E_b = P_t/R_t = T_c P_t/R. \tag{21}$$

If the terminal's transmitter has the same output power regardless of the value of $T_c$, that is, the terminal is peak-power limited; it transmits the same instantaneous power regardless of its duty cycle. Note from (21) that the energy transmitted per bit of data is proportional to $T_c$; that is, reducing the duty cycle reduces $E_b$. Reduction in $E_b$ reduces the S/N ratio of signals received from the terminal by the satellite and increases the effectiveness of a jammer proportionately. In other words, TDMA operation with a peak-power-limited HPA reduces antijam capability.

### 3.4.2.2 Average Power Limited

If the terminal's HPA has the fundamental characteristic that

$$P_t = P_a/T_c. \tag{22}$$

where $P_a$ is the HPA's average power output, its antijam capability is independent of $T_c$, the TDMA duty cycle. However, $P_t$ can be substantially larger than $P_a$ perhaps stressing the associated RF circuit elements to, or above, their limit of operation and/or exceeding the maximum power output capability of the HPA.

When $P_t$ is given by (22) the HPA is said to be average power limited. Substituting (22) into (21) gives

$$E_b = P_a/R. \tag{23}$$

demonstrating that $E_b$ is independent of $T_c$. This independence, in turn, implies the terminal's antijam capability is independent of its TDMA duty cycle.

Average-power-limited TWTA HPAs are, in general, more difficult to build, have a shorter life, and are more costly than peak-power-limited TWTA HPAs. It is also true that average-power TWTAs seldom have a peak power more than 10 times its average-power output. Radar HPAs are an exception to this rule because they have extremely short "on" time and a very low duty cycle. The short "on" time is essential because it prevents the HPA from reaching thermal equilibrium; that is, its thermal time constant is longer than the time duration of the transmitted pulse, or burst of data. Current state-of-the-art TWTAs can be designed to operate as an average-power-limited HPA with duty cycle greater than 0.1. The associated thermal time constant is about 100 microseconds. For data

bursts lasting more than 100 microseconds the HPA is peak-power limited. For data bursts between 10 and 100 microseconds long the HPA is average-power limited.

### 3.4.3 <u>Signal Processing Considerations</u>

Signal processing characteristics should be considered whenever they interact with antenna and antijam waveform design. If the latter is some form of frequency hopping, the hop rate is an important and interactive signal processing characteristic. It will be necessary to demodulate the uplink signals and remodulate them for downlink transmission if the uplink access mode is different from the downlink access mode. Some requirements are well known; yet, it may be beneficial to repeat them here so as to emphasize their importance. The following sections address these signal-processing characteristics briefly.

#### 3.4.3.1 Hop Rate

The principal antijam characteristic, of a frequency-hopping (FH) spread-spectrum waveform, is the inability of an interceptor, or jammer, to determine a MILSATCOM system's instantaneous operating frequency band. If the users dwell on a frequency long enough, a jammer can determine it and concentrate all of its jamming signals in the known frequency band. If a jammer can do this continuously or consistently, the antijam waveform is essentially ineffective. Reducing the dwell time $T_h$ makes it more difficult for the jammer to determine the operating frequency in sufficient time to effectively compromise the waveform's antijam characteristic. Increasing the hop rate above the data rate requires reconstruction of the data bits through integration of several chips resulting in a noncoherent combining loss and/or using an appropriate code rate. Consequently, there is an optimum hop rate.

A jammer designed to defeat a FH antijam waveform must determine the user's instantaneous operating frequency, and set its transmit frequency equal to it in less time than a user remains at that frequency. Using state-of-the-art devices, a jammer can determine the user's instantaneous frequency and tune its transmitter to it in less than a microsecond. However, the path length $L_u$ between the user and the satellite is always less than path length $L_j$ from the satellite to the jammer plus the path length $L_{ju}$ from the jammer to the user. Whenever the differential delay $T_d$ is larger than the dwell time $T_h$, the jammer is essentially ineffective. For known, or expected scenarios, it is possible to eliminate the frequency follower jammer (FFJ) as a threat by choosing the hop rate $F_h$ such that

$$F_h \geq c/(L_j+L_{ju}-L_u), \tag{24}$$

where c equals the velocity of light.

For a worst-case scenario the differential path length $DL = L_j+L_{ju}-L_u$ is given by

$$DL = L_{ju}(1-\cos A), \tag{25}$$

where A is the elevation angle of the satellite measured at the user terminal. It follows that if the hop rate is equal to or greater than $c/L_{ju}(1-\cos A)$, the jammer's signals will arrive at the user, for a downlink jammer, or at the satellite, for an uplink jammer, after the user has hopped to a new frequency. For A = 20 degrees and $L_{ju}$ = 100 miles, $F_h$ should equal 28 khps. Choosing $F_h$ larger than 28 khps will unnecessarily compromise the user's communication S/N ratio. Choosing $F_h$ less than 28 khps will increase both the effectiveness of the jammer and the effective signal power received by the user. The latter increase is substantially less than the former; hence the "optimum" hop rate for this case is approximately

25 khps. It is important to note the assumed jammer would need to be at least three miles above the earth's surface in order to have a line-of-sight path to the terminal. This forces the jammer to be airborne and substantially limits its capability compared to a ground-based jammer.

#### 3.4.3.2 Demodulate/Remodulate

Demodulation of the uplink signals permits the satellite to establish a BER on the uplink that depends on the uplink S/N ratio, signaling format, error code detection method, etc. Using the results of this signal processing to generate a bit stream for modulation of the downlink tends to isolate the uplink BER from the downlink BER.

If users operate in a FDMA mode on the uplink and a time division multiplex (TDM) mode on the downlink, it is necessary to demodulate the uplink signals to obtain the bit stream and then use it to modulate the downlink. This format (i.e., FDMA on the uplink and TDM on the downlink) permits maximum uplink antijam and maximum downlink "transmission efficiency." The latter is explained in the next section.

#### 3.4.3.3 Requirements

Whenever multiple users access a satellite they have the potential to "jam" one another in any one of three principal ways. These are discussed in the next sections.

##### 3.4.3.3.1 Small Signal Suppression

When at least one signal is much larger than all other signals passing through the same HPA, the smaller signals are suppressed up to 6 dB. This small signal-suppression phenomena can be reduced substantially when the satellite demodulates the uplink and modulates the downlink with the resultant bit

streams. This assumes the satellite's on-board processor normalizes the relative signal strength of the bit streams so the signal strengths of the bit streams are nearly equal.

3.4.3.3.2 Multiple Carrier Intermodulation

Whenever two or more modulated carrier signals f1, f2, ...fn are impressed on a nonlinearity, the output signals contain intermodulation signals at $f1 \pm f2$, $f1 \pm f3$, etc. These intermodulation signals can fall in incorrect frequency bands and "jam" the desired signals. Modulating a single carrier with several bit streams does not tend to result in this intermodu- lation noise problem. If the multiple bit streams are time-division-multiplexed and modulate a single carrier, intermodulation noise is completely eliminated.

3.4.3.3.3 Power Robbing

A perfectly linear HPA shares its output power in proportion to the relative signal strength of the input signals. Consequently, a repeating (i.e., no demod/remod) satellite tends to allocate more downlink EIRP for stronger uplink signals than for weak uplink signals. This can happen because a user is not properly disciplined; that is the user transmits with more EIRP than assigned. This lack of discipline deprives the disciplined user from its fair share of the downlink power. This process is referred to as "power robbing." If the satellite demodulates the uplink, and normalizes the resultant bit stream, power robbing is eliminated.

3.5 ESTIMATED WEIGHT AND POWER

Weight and power estimates are accurate only after a design is completed. Weight and power of initial concepts like those presented here are not easily estimated accurately. Since the candidate antenna is similar to that described by

Dr. Mayhan of Lincoln Laboratory, the estimates given by him are used to estimate the weight and power required by the candidate antenna. However, the candidate system has 285 beams and a single lens versus 271 beams and four lenses proposed by Mayhan. These changes are included in Table 1-3. Also the additional weight and power are shown if a ferrite-switching isolator is connected to each beam port to provide better than 40 dB of all signals incident in the "off" beams.

### 3.5.1 Payload Weight and Power

Table 1-3 summarizes the estimated payload weight and power requirements. Without consideration for redundancy the antenna system will weigh approximately 110 lbs and require about 110 watts. This ripples through the spacecraft requiring an estimated 300-lb launch weight (i.e., weight approximately 1.7 payload weight plus .6 payload power in watts) and 110 watts launch power.

### 3.5.2 Redundancy Estimate

Redundancy requirements depend on a reliability analysis to first indicate those single-thread points that can and should be made redundant. This in turn requires a reasonably detailed design. However, using the same rough order of magnitude method that estimated the launch weight, the launch weight should increase about 50 lbs to account for redundant components. Allowing an additional 20-percent contingency results in an estimated launch weight of 420 lbs and 110 watts.

### 3.5.3 Integration on DSCS III

The candidate design will require a volume approximately 30" in diameter and 45" long. If a transmit MBA or the GDA could be removed there would be ample space for the candidate EHF antenna and its related SHF (20 GHz) transmitting antenna.

This EHF antenna requires substantial contiguous space, a rare commodity on the DSCS III spacecraft. Further consideration of integration should follow a more detailed design of the EHF antenna and the associated SHF transmitting subsystem. Although the latter was not a subject of this study it is considered briefly in the next section.

### 3.5.4 Downlink Antennas

The need to provide switched-beam high-EIRP service to the EHF users indicates either an MBA with a beam-switching network driven by a TWTA, or an active aperture antenna. Because of the similarity between the MILSTAR downlink requirements and the EHF/Wideband user requirements, it is recommended the MILSTAR downlinks MBA and TWTA be used to support the uplink EHF antenna subsystem. Alternatively the GDA could be converted to a dual-frequency device and used to support high-data-rate communication on the downlink to the continental (CONUS); an active aperture antenna could be added to support EHF user downlinks to the remote sites. The latter need not occupy a space more than 6" x 6" x 12". Further details are beyond the scope of this report.

### 3.5.5 Frequency Reuse Considerations

Chapter 2 of this report addresses the issues of a dual-polarized uplink antenna system operating at SHF. The general statements and requirements discussed apply equally to frequency reuse on the EHF uplink. Depolarization due to propagation through rain is more pronounced and it will be, in general, more difficult to build devices capable of maintaining or adapting to the desired polarization. In short, the prospect of frequency reuse at SHF is at best very diffult. Satisfactory frequency reuse at EHF will be even much more difficult to achieve because of the shorter operating wavelength and concomitant required dimensional accuracy.

Adequate consideration of this subject is beyond the scope of this report; however, frequency reuse should be considered in detail once a single polarization payload design is formulated.

UNCLASSIFIED

SECURITY CLASSIFICATION OF THIS PAGE

AD-A162 114

# REPORT DOCUMENTATION PAGE

**1a. REPORT SECURITY CLASSIFICATION**
UNCLASSIFIED

**1b. RESTRICTIVE MARKINGS**

**2a. SECURITY CLASSIFICATION AUTHORITY**
N/A

**2b. DECLASSIFICATION/DOWNGRADING SCHEDULE**
N/A

**3. DISTRIBUTION/AVAILABILITY OF REPORT**
Unlimited

**4. PERFORMING ORGANIZATION REPORT NUMBER(S)**
MSO-85-114, Final Report, ~~Volume~~ Part II

**5. MONITORING ORGANIZATION REPORT NUMBER(S)**

**6a. NAME OF PERFORMING ORGANIZATION**
M/A-COM LINKABIT, Inc.

**6b. OFFICE SYMBOL** (If applicable)

**7a. NAME OF MONITORING ORGANIZATION**
Defense Communications Agency
MILSATCOM Systems Office, Code A800

**6c. ADDRESS** (City, State and ZIP Code)
8619 Westwood Center Drive
Vienna, VA 22180

**7b. ADDRESS** (City, State and ZIP Code)
Washington, D.C. 20305-2000

**8a. NAME OF FUNDING/SPONSORING ORGANIZATION**
Defense Communications Agency

**8b. OFFICE SYMBOL** (If applicable)
A800

**9. PROCUREMENT INSTRUMENT IDENTIFICATION NUMBER**

**8c. ADDRESS** (City, State and ZIP Code)
8th & S. Courthouse Road
Arlington, VA 22204

**10. SOURCE OF FUNDING NOS.**

| PROGRAM ELEMENT NO. | PROJECT NO. | TASK NO. | WORK UNIT NO. |
|---|---|---|---|
| | | | |

**11. TITLE** (Include Security Classification)
(see Abstract section)

**12. PERSONAL AUTHOR(S)**
L. Ricardi

**13a. TYPE OF REPORT**
Final

**13b. TIME COVERED**
FROM 4/30/85 TO 11/30/85

**14. DATE OF REPORT** (Yr., Mo., Day)
85 November 29

**15. PAGE COUNT**
135

**16. SUPPLEMENTARY NOTATION**

**17. COSATI CODES**

| FIELD | GROUP | SUB GR |
|---|---|---|
| | | |
| | | |

**18. SUBJECT TERMS** (Continue on reverse if necessary and identify by block number)

**19. ABSTRACT** (Continue on reverse if necessary and identify by block number)


11. DSCS/Wideband SHF Enhancements and EHF Antenna Recommendations for the Mid- (1986-1991) and Far-Term (1992-2000) Periods (U), Volumes I and II

(U) The primary purpose of this effort is to examine SHF modifications to DSCS/Wideband systems for all users in the mid- (1986-1991) and far-term period (1992-2000). This report consists of two parts.

Part I summarizes the recommendation for further study provided by the SHF Working Group during the July through November 1985 period. Part I also includes preliminary examination of the benefits of high power amplifier linearization techniques.

Part II of this document provides detailed analysis and recommendations on the implementation and potential benefits of SHF frequency reuse through polarization diversity.


**20. DISTRIBUTION/AVAILABILITY OF ABSTRACT**
UNCLASSIFIED/UNLIMITED ☒ SAME AS RPT. ☐ DTIC USERS ☐

**21. ABSTRACT SECURITY CLASSIFICATION**
UNCLASSIFIED

**22a. NAME OF RESPONSIBLE INDIVIDUAL**
Lt Col. R. Smart

**22b. TELEPHONE NUMBER** (Include Area Code)
(202) 692-1949

**22c. OFFICE SYMBOL**
A800

DD FORM 1473, 83 APR

EDITION OF 1 JAN 73 IS OBSOLETE.

UNCLASSIFIED

SECURITY CLASSIFICATION OF THIS PAGE

END

FILMED

1-86

DTIC